TECHNICAL EVALUATION OF A RADIO DIRECTION FINDING SYSTEM UTILIZING COMPLEMENTARY SE-QUENCE PHASE CODING OF INCIDENT SIGNALS

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THESIS

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by

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June 1973

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Submitted in partial fulfillment of the
requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

from the

NAVAL POSTGRADUATE SCHOOL
June 1973

ABSTRACT

A new radio direction finding system has been proposed which offers advantages over current systems in the areas of bearing accuracy and simplicity of antenna design requirements. This system phase codes the outputs of a linear array of omnidirectional antennas, and then combines these time-separated signals so that the sequences produced uniquely determine the signal arrival angle. A correlation process using matched filters is performed on the resulting sequences in order to determine the corresponding arrival angle.

Resolution of the inherent ambiguities of the system, due to elevation angle and adjacent quadrant azimuth components is demonstrated using a two-array technique. A computer simulation of the phase coding/decoding process, using an idealized signal model is used to evaluate the performance of the system for various input noise conditions, array lengths, and input signal parameters.

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ACKNOWLEDGEMENT

The author is indebted to Professor S. Jauregui for suggesting the area of this thesis. Additionally it was Dr. Jauregui's own research and that of two of his thesis students, C. E. French and R. C. Todaro, which provided much of the necessary background material.

I. INTRODUCTION

A. STATEMENT OF THE PROBLEM

In order to determine the geographical location of a radio frequency-emitter, its emissions must be detected and processed simultaneously by two or more radio direction finding (RDF) systems. The unambiguous determination of the angle of arrival (AOA) of RF energy received from a target transmitter enables each RDF system to generate a line-of-bearing along the Poynting vector associated with the transmitted energy. Ideally, the lines-of-bearing can then be plotted on a chart containing the known locations of the RDF sites, and the point of intersection, or the area enclosed by three or more lines-of-bearing, will indicate the approximate transmitter location.

The accuracy with which a target transmitter can be located by such an operation depends upon many factors. The number and precision of the individual AOA measurements, and the relative positions of the transmitter and RDF sites are probably the most important; and each of these factors is, in turn, dependent upon numerous other considerations, both technical and non-technical. The point of intersection of two near-parallel lines-of-bearing (LOB) will vary significantly with relatively small standard deviation errors in either LOB, while the intersection of two near-perpendicular lines-of-bearing will be influenced only slightly by the same errors. It is obvious that additional

AOA data, providing a wide dispersion of available lines-of-bearing, can greatly enhance emitter location accuracy over a large geographical area.

Increasing the number and geographical dispersion of radio direction finding sites has proven very effective in the improvement of emitter location in the HF communications spectrum (2 - 30 MHz). The problem of accurate AOA measurements is particularly acute in this frequency range since most signals of interest are received via ionospheric propagation. The long propagation paths involved, and the non-deterministic nature of ionospheric propagation, have necessitated the use of probabilistic techniques applied to multiple RDF measurements. The availability of multiple LOB inputs from RDF sites in a coordinated net allows a central processor to sort through the inputs using an optimization scheme, and to choose those inputs which yield a most probable emitter location with a minimum area of ambiguity.

From a technical standpoint, the greater the number and dispersion of RDF sites, the larger the statistical data base from which decisions can be made. However, from economic and political standpoints, the number and placement of complex, large-array RDF systems for operation in the HF range are very limited. The development of a simpler, more compact HFDF system would allow the strategic placement of additional sites in those locations where more complex facilities do not exist, or are impractical.

B. STATEMENT OF PURPOSE

The purpose of this paper is to investigate the feasibility of a new and unique radio direction finding concept, and to determine the applicability of this concept to a realizable system for operation in the HF communications range. The fundamental concept was originally presented in the Masters thesis of R. C. Todaro [1]. The results of this original work indicated that the new system concept offered definite advantages over current RDF systems in the areas of bearing accuracy and simplicity of antenna design.

The specific areas of investigation described in this paper are: (1) a discussion and analysis of the ambiguities inherent to the system, (2) a qualitative analysis of system performance in the presence of background noise; and, (3) physical scaling of the system for operation in the HF frequency range.

The remainder of this chapter contains a brief description of a generalized RDF system, and a comparison of the RDF systems currently in use. Chapter Two contains a description of the new system, and a discussion of geometric ambiguities. Chapter Three describes the computer model of the system and the results obtained from the incorporation of background noise and scaling parameters.

C. GENERALIZED RDF SYSTEM

Before examining any radio direction finding system for its individual characteristics, it is useful to consider the

elements basic to any system and understand how they can affect its performance. A block diagram of a generalized RDF system is shown in Figure 1-1.

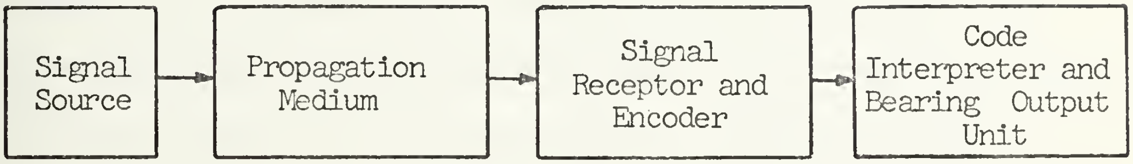


Figure 1-1. A GENERALIZED RDF SYSTEM

1. <u>Signal Source</u>

The signal source has only three properties which will affect the operation of an RDF system. These are its placement on the earth, its transmission frequency, and its type of modulation. Its location on the surface of the earth will determine the great circle bearing between the transmitter and the RDF location, but except in very simple cases this is insufficient information with which to predict the line of bearing readout from the RDF system. The transmission frequency will affect the results obtained by determining, in conjunction with the transmission medium, the mode by which the signal is sent from the source to the RDF site. The type of modulation will have an effect on those systems which compare arrival times of specific modulation events, or those which integrate multiple antenna sweeps over each line of bearing.

2. Propagation Medium

It is possible for the propagation medium to have little effect on the angle of arrival of the received signal, but this is not normally the case. If the signal arrives at the RDF site by ground wave propagation, then results are relatively simple and easily predicted. Most of the signals of interest arrive via the ionoxphere, however, so its effects must be taken into account. There has been a great deal of work done to investigate the character of the ionosphere, and much more remains before accurate predictions of the transmission path of a received signal can be obtained. A large part of the difficulty lies in the complexity of the equipment necessary for measurements on the ionosphere, and the large amount of time necessary to complete a set of measurements. Since the characteristics of the ionosphere depend on the sunspot cycle as well as seasonal and diurnal changes, a complete set of measurements on a given path could not be obtained in less than eleven years.

If accurate transmission path prediction were available it is conceivable that the emitter location process could be performed by a single-site RDF system. That is, azimuth and elevation AOA measurements could be used in conjunction with path prediction to compute the point of origin of received signals. Such a capability would greatly enhance the emitter location capability of a coordinated HFDF net.

3. Signal Receptor and Encoder

This is the first section of an RDF system over which some design control can be exercised, and the choice of this block will determine to a great extent the accuracy and speed with which a line-of-bearing can be generated from the received signal. The basic purpose of this unit is to sense the incoming signal, combine the proper components to give an output which can be uniquely interpreted as an angle of arrival, and present these components to a code interpreter for transformation to a form suitable for display. Some of the common types of receptors and encoders will be discussed in a later section describing the RDF systems currently in use.

4. Code Interpreter and Bearing Output Unit

In most RDF systems the functions of this section are performed jointly by an electronic indicator sub-system and the human operator. Ideally, the code interpreter should be carefully matched to the characteristics of the coder so the desired output information can be obtained in the shortest time and with the highest possible accuracy. There is, of course, no way in which the equipment associated with these functions can increase the fundamental accuracy of the system to a greater degree than is allowed by the components of the receptor and encoder, but it should not degrade their performance. An exact match occurs between the signal encoder section and the code interpreter of the proposed system through the use of a matched filter technique.

That is, the encoding section generates a phase coded sequence which is uniquely determined by the angle of arrival of the incoming signal. The inverse filter in the code interpreter which provides an exact match for this sequence will produce a unique autocorrelation (or cross correlation) output. This process is described in more detail in the following chapter.

D. CURRENT RDF SYSTEMS

There are four basic system types which are commonly used to determine the AOA of transmitted signals. The choice of which system to use in any particular application is determined primarily by the frequency range of interest, the degree of accuracy desired, and the type of signal modulation expected. These four basic types are as follows:

1) Relative amplitude
2) Relative phase
3) Time of Arrival (TOA)
4) Doppler

1. Relative Amplitude Methods

By far the most commonly used RDF systems, these employ directional antennas whose patterns may be mechanically or electronically rotated. The line of bearing is then determined by observing the angular position of the antenna (or its beam pattern) when the system response is either at a maximum or at a minimum value. Those systems which employ physically small antennas (e.g., magnetic-field

sensitive loops) mechanically rotate the antenna while observing the system output.

A loop antenna which is very much smaller than a wavelength has two sharp nulls in its pattern, and these nulls lie in a plane normal to the plane of the loop. Rotating an antenna of this type and noting the positions of the nulls will give an indication of the bearing of the incoming signal. The bidirectional ambiguity can be resolved by addition of the signal picked up by a separate omni-directional antenna located closely to the loop, but not so close as to affect the pattern of the loop. Addition of the antenna outputs will produce a cardioid pattern indicating the proper null to use for the bearing.

The signal processing portion of this system (code interpreter and bearing output) consists very simply of a standard communications receiver whose output feeds the deflection circuits of a circularly swept CRT display. Antenna position is synchronized to the CRT sweep to establish a common rotation rate and a fixed reference bearing.

The primary disadvantage of this system is the relatively low overall gain of the antenna, and the fact that the distinct feature of the pattern (null) occurs where the gain is at a minimum. For these reasons the loop antenna is ineffective in a low signal level/high noise environment. In addition, the loop antenna is subject to polarization errors. For ionospherically propagated waves,

which have nearly random polarization, these errors may be as large as ±35° making the loop antenna worthless for long range DF in the high frequency band [2].

Instead of using a rotating element, two or more fixed elements may be installed with their respective nulls symmetrically spaced. These individual elements may be loops, pairs of monopoles (thereby minimizing the polarization error mentioned above), or any structure which will produce a "figure-eight" pattern with distinct nulls. The outputs from these elements are combined in a special motor/-transformer in which they are connected to symmetrically spaced stator windings. A rotating pickup coil (rotor) acts as the secondary winding, and its output is fed to the RDF receiver. The amplitude output of the receiver can then be compared with the mechanical position of the rotor of the transformer, or goniometer, and the bearing of the signal obtained. Bidirectional ambiguities can be resolved by forming a cardioid pattern as in the rotating loop system.

While the figure eight patterns of the individual elements have relatively broad beamwidths, the combining action of the goniometer produces a sharper "propellor" pattern with a beamwidth inversely proportional to the number of individual element pairs. This allows the use of the maximum gain portion of the array pattern for determination of AOA. However, in order to achieve good figure eight patterns, small aperture elements (much less than a wavelength) must be employed. Therefore the array still

has a low overall gain characteristic. In addition, the small aperture requirement severely limits the number of non-interacting antenna elements which may be used to comprise the array. This limits the factor by which the overall antenna pattern beam width may be reduced, and, in turn, limits the angular resolving capability of the array.

In order to improve the gain and bearing resolving capabilities of relative amplitude RDF systems, large aperture arrays have been developed. These arrays are usually composed of a large number of individual antennas located on the periphery of a symmetrical structure, ordinarily a circle, and connected to a goniometer which feeds a single channel receiver. The code interpreter/bearing output section of this system is similar in design and operation to that associated with the simple spinning loop system.

The large aperture RDF arrays that are most commonly in use today are of the Wullenweber design. This is a multiband circular array, each section of which contains an independent concentric array of vertical monopoles, and a circular reflecting screen. Two or three concentric sections, each with different numbers and sizes of active monopole elements, are designed to cover overlapping portions of the HF communications spectrum. The outputs from each of the individual antenna elements are fed to a capacitive coupled goniometer, which utilizes discrete rotor and stator pickup plates instead of inductively linked coil windings.

The use of capacitive coupling, required by the large number of individual antenna elements, enables the code interpretor/bearing output section to take full advantage of the fine-grain resolving capability of the antenna array. The goniometer forms a very narrow array beam from a set of 40 pickup plates located on the rotor, with these 40 plates feeding delay lines that act to compensate for the physical structure of the array and transform it into an equivalent linear (broadside) array.

The Wullenweber can thus produce a very narrow rotatable beam providing resolving power of approximately 3°. However, in order to achieve this resolving power, up to 120 separate antenna elements are required for each frequency range, and the physical dimensions of the array commonly exceed 1000 feet in diameter and 200 feet in height. Therefore, the array and its associated transmission line system is a very complex and expensive structure.

All relative amplitude systems, regardless of array size, have the disadvantage that they are not instantaneous. That is, the amount of time the narrow array beam spends at each azimuth angle is a small fraction of its total rotational period. This can produce erroneous and misleading results with signals of short duration such as on/off keyed pulse transmissions. The scan rate may be increased to overcome these errors, but a corresponding increase in the RDF receiver bandwidth is required in order to recover the rapidly sampled signals. Therefore, the maximum scan rate

is limited by the receiver bandwidth necessary for good co-channel discrimination (selectivity).

2. <u>Relative Phase Methods</u>

These systems utilize the measured signal phase difference between omnidirectional antennas to determine angle of arrival. This phase difference is given by the equation:

$$\phi = 2\pi\frac{d}{\lambda}\cos\theta \quad \text{(radians)}$$

where,

d = antenna separation along the baseline

λ = wavelength of incident wavefront

θ = angle of arrival, measured from the baseline.

It is obvious that, assuming a given accuracy in phase measurements, the accuracy with which the AOA can be determined (i.e., the resolving power of the system) is greater for larger values of the ratio $\frac{d}{\lambda}$. When $\frac{d}{\lambda}$ is greater than unity, however, there is a possible ambiguity of 2π, or a multiple of 2π, in the measured phase angle, and there will not be a unique solution for the AOA.

Providing that $\frac{d}{\lambda}$ is not too large however, the alternative solutions are well separated, and it is generally possible from other considerations to select the most probable [3]. For example, the approximate wavelength, or range of wavelengths, of the incident signals can be determined from the tuned frequency of the RDF receivers,

le the separation (d) is a fixed constant. Simple logic cuitry can be employed to associate a range of values of phase shift with corresponding frequency bands of the receivers.

An additional ambiguity results when the incident nal wavefront has a non-zero elevation angle component to ionospheric propagation. The phase difference ation will contain an additional cosine factor of the vation AOA, and the one-equation/one-unknown situation l no longer apply. The requirement for an additional sured parameter to uniquely determine both elevation and muth angles of arrival necessitates the use of a second n-collinear) antenna pair. The second antenna pair is ally placed orthogonally to the first, resulting in a xially symmetric array. Since the elevation angle ponent must be considered in the proposed new system, s topic will be discussed in more detail in Chapter II.

The signal processing portion of the relative phase tem utilizes two receivers, one for each antenna pair. IF outputs of these receivers are feed to a sum and ference unit which accurately adds and subtracts the nals and rotates the phase of either the sum or the ference by $\pi/2$. The resultant sum/difference signals n drive orthogonal deflection plates of a CRT display t, thus achieving the effect of a goniometer rotating the intermediate frequency of the receivers. The display tern produced by this technique is an ellipse with the

azimuthal line of bearing along its major axis. This signal processing scheme produces cancellation of the effects due to elevation angle of arrival.

A system of this type is difficult to build and maintain because of the extreme requirements on the performance of the receivers. Both channels must be phase and gain matched over the operating range in order to maintain the amplitude and phase relationship between the received signals. In addition, the frequency range that can be covered with a given antenna spacing is limited. The receiving, amplifying, and phase comparison portions of the system can be designed to cover a wide frequency range. But the antenna system cannot in practice be made to cover a range wider than about four to one in frequency if adequate resolving power without too many ambiguities is to be obtained [3].

3. Time of Arrival Methods

The time of arrival (TOA) technique requires the measurement of a given modulation event at two known antenna locations. This modulation event may be the crest of a sinusoidal modulation signal, or a specific point (leading or trailing edge) of a pulse signal. This technique is frequency independent and, thus, does not experience the wavelength/antenna separation ambiguities inherent to the relative phase technique. However, elevation angle components produce ambiguities just as in the relative phase technique.

The TOA method is analogous to the relative phase method in that antenna baselines are established, and gain/-phase matched receivers must be used in the signal processing section.

4. Doppler Methods

The Doppler principle of wave motion can be applied to a specially constructed antenna system to determine the angle of arrival of a signal [4]. If an omnidirectional antenna is rotated at a fixed rate in a circular path, the signal at the antenna output will be a frequency modulated representation of the incident signal. The peak frequency deviation of this signal will be:

$$\Delta f = \frac{R\omega}{c}$$

where

C is the speed of propagation (meters/sec),

R is the path radius (meters), and

ω is the rotation rate (radians/sec.)

The maximum frequency deviations will occur when the antenna is moving directly away (negative Δf) or directly toward (positive Δf) the emitter. Although this property is not affected by elevation angle components of the incident wavefront, the peak deviation frequency will be diminished by the cosine of the elevation angle and may reduce the resolution capability of the system.

In practice, the antenna would not be physically rotated; instead a ring of omnidirectional antennas would be constructed and a capacitively coupled goniometer would sample each element. The main difference between this technique and the Wullenweber system is that the Doppler technique samples individual low gain elements while the Wullenweber forms the equivalent high gain beam pattern of a 40-element broadside array.

The primary disadvantage of this system is that it requires a circular array several wavelengths in diameter, thus precluding its use in the lower end of the HF frequency range.

II. A NEW RDF SYSTEM

A. BASIC DESCRIPTION

A new direction finding technique has been proposed which utilizes the finite difference between arrival times of distinct modulation events as an incident RF plane wavefront propagates along a linear antenna array. Although the time of arrival concept is not new, the technique by which signals are processed, coded, and interpreted to extract angle of arrival information is considered to be unique. [1]

Basically, the technique involves the binary phase coding and correlation of the time separated antenna outputs. The correlation of these coded signals in matched filter devices produces a time-compression of the signals, causing them to arrive at the filter output at the same point in time. When this occurs, a unique correlation output is produced. Depending on the choice of phase codes utilized, this unique output may be either an amplified replica of the original input signal (autocorrelation), or it may be a distinct null (crosscorrelation).

The characteristic of this process which enables the determination of arrival angle is the one-to-one relationship between AOA and the particular matched filter configuration which produces this distinquishable output. In practice, the outputs form a bank of matched filters, constructed to

match a range of arrival angles, would be scanned for this unique output. The filter producing that output would correspond directly to a particular arrival angle.

The technique outlined above makes use of a special class of binary codes called complementary sequences. The time correlation of these sequences in matched filters will produce a unique output which is free of the coding noise prevalent with all other discrete binary codes. Before describing in detail the application of these codes to the RDF system, however, a brief discussion of general phase coding techniques is considered necessary for continuity. Therefore, these techniques, and the specific advantages of complementary sequences, are presented in the following section.

B. PHASE CODING TECHNIQUES

Phase coding techniques were first used (in communications systems) to improve the range capability of pulse radars. While not relevant to the topic of this paper, this utilization of phase coding provides a graphic description of general coding techniques and an example of their application to an actual system.

Two parameters of a transmitted radar signal may be varied in order to increase the total RF energy illuminating a radar target. These are the peak pulse power, and the pulse duration. The peak power of the transmitted pulses may be increased by utilizing higher power amplifiers in

the transmitter output stage. However, this has proven inefficient from a power standpoint, since many radar transmitter tubes are peak power limited while operating at only a small fraction of their average power capabilities. The average power output of these tubes may be increased by widening the transmitted pulses. However, this too has a drawback because the range resolution capability of a pulse radar is inversely proportional to pulse width.

As a result of this dilemma, intrapulse modulation techniques were developed so that pulse widths could be increased without sacrificing range resolution. The most common example of intrapulse modulation is the "chirping" or pulse compression system in which the RF carrier frequency of each pulse is swept between fixed limits during its transmission. The received pulses are then passed through a matched filter whose response characteristic (delay versus frequency) is the exact inverse of the transmitted pulse characteristic (frequency versus time). This allows all the frequency components of the wide reflected pulse to arrive at the output simultaneously, resulting in a much narrower, amplified pulse signal to the detector stage. Phase coding of the transmitted pulse is another technique of intrapulse modulation.

The phase coding technique utilizes a discrete process to encode pulses so that each return pulse, when input to a matched filter, causes discrete, coded segments of the pulse to arrive at the output simultaneously. Each radar pulse is

divided into "n" equal time segments, and a coded sequence of "phase states" is assigned to these segments. The states determine the phase relationship between the RF carrier during that segment, and a fixed phase carrier reference. If the only allowable phase states in each segment are 0° or 180°, then the coding is termed binary phase coding. Although multiphase coding schemes have been devised and used in certain applications, only binary phase coding will be discussed here.

A representation of a binary phase coded pulse is shown in Figure 2-1(a). The RF carrier is divided into seven equal time segments, with each segment assigned a plus (+) or minus (-) phase state. The phase state values are here defined as no carrier phase reversal (+) and a 180° carrier phase reversal (-). A unit amplitude is assumed for the pulse.

The filter which provides an exact match for this coded pulse is shown in Figure 2-1(b), where the discrete delay units are equal to the pulse segment durations and the (+) and (-) phase reversals occur in reverse order from the transmitted sequence phase states. In a physically realizable filter, the discrete time delays would be produced by delay lines or electronic delay circuits, and the phase control would be provided by buffer amplifiers with inverting or non-inverting outputs. For this example, the buffer/phase control devices are assumed to have unity voltage gains.

Figure 2-2 is a time diagram showing the output waveforms at the intermediate points "a" through "g", and the final summation output waveform.

There are two characteristics to be noted about this output. First, it contains a narrow main pulse that is delayed n-1 time units from the leading edge of the received pulse. Since this narrow, main pulse would be utilized by the radar receiver's detector and timing circuits to determine target range, this delay must be compensated for. Second, there are undesired secondary pulses in the output waveform which might appear to the radar receiver as excess noise or false targets.

There are other codes, called pseudo-random noise codes, which will produce smaller secondary pulses than those produced by the (randomly selected) code utilized in this example. However, all single-sequence coding schemes of this type produce a certain amount of "coding noise", as these secondary pulses are called, and therefore exhibit a finite processing signal-to-noise ratio.

There is a method, however, which will completely eliminate the coding noise, although it requires the simultaneous transmission of two distinct coded sequences. These codes, called complementary sequences, were first utilized by M. J. E. Golay in the field of multislit spectrometry. Since then, they have been utilized in various radar and sonar system designs. Golay mentioned in 1961 that these

sequences might be applied to data transmission [5], and there has been some work toward that end [6].

A great deal of work has gone into the investigation of properties of complementary sequences, and the methods by which they may be generated. However, only the basic characteristics of these sequences will be presented here. For more detailed studies, the reader is referred to References 5 and 7 listed at the end of this paper.

A complementary sequence pair is defined to be a pair of equally long binary sequences which have the property that the number of pairs of like digits (or elements) with any given separation in either sequence is equal to the number of pairs of unlike elements with the same separation in the other sequence.

For example, considering the following sequences

A: - - - + - - + -

B: - - - + + + - +

the number of like elements in A with separation distances of one element is three:

A: - - - + - - + -
l l u u l u u ..... 3 like elements

The number of unlike elements in B with separation distances of one element is also three:

B: - - - + + + - +
l l u l l u u ..... 3 unlike elements

Similarly, the number of like elements in A and the number of unlike elements in B for all separation distances are listed below:

Separation Distance	# Likes in A	# Unlikes in B
2	3	3
3	4	4
4	2	2
5	2	2
6	1	1
7	1	1

Golay has shown [5] that the sequence lengths must always be an even number, and must be the sum of two squares. For example, the possible sequence lengths up to 50 are: 2, 3, 8, 10, 16, 18, 20, 26, 34, 36, 40, and 50. He has also shown [5] that 2^6 complementary sequence pairs (some identical) can be generated by performing the following basic operations on a given sequence pair:

1. Interchanging the sequences
2. Reversing the order of the first sequence
3. Reversing the order of the second sequence
4. Complementing the elements of the first sequence
5. Complementing the elements of the second sequence
6. Complementing the even-order elements of both sequences

An example of a pair of complementary sequence coded pulses being processed by a matched filter pair is shown in Figures 2-3 and 2-4. The input pulses, utilizing complementary sequence codes of length eight, are shown in Figure 2-3(a). The pulses and their matched filters, shown in

Figure 2-3(b), are labeled 'A' and 'B' respectively. The outputs from these matched filters, and the final combined output, are shown in Figure 2-4. This final output waveform shows the signal processing advantage of this coding scheme. The total energy content of both received pulses has been compressed and combined into a single narrow pulse with an amplitude gain factor of twice the code length. In addition, the complete cancellation of all undesired coding noise indicates that an infinite processing signal-to-noise ratio can be obtained with this technique.

In radar system applications, this property allows transmitters to be operated at high average power levels while utilizing pulses with high frequency content. Thus, both transmitted energy and range resolution goals may be met. The price that must be paid for this advantage is the increased complexity of the signal generation, transmission, and detection components.

The signal coding/decoding process illustrated above may also be described in mathematical terms. If we let a_i and b_i be the elements of <u>any</u> pair of n-element binary sequences (taking on values +1 or -1), and let a_i' and b_i' be the phase reversal portions of their corresponding filters, then the output sequences from the filters can be represented by:

$$c_j = \sum_{i=1}^{n-j} a_i a_{i+j}'$$

$$d_j = \sum_{i=1}^{n-j} b_i b_{i+j}'$$

where the subscript "j" takes on integer values from -n+1 to n-1 and, by definition:

$$c_{-j} = c_j$$

$$d_{-j} = d_j$$

The summation equations represent a discrete convolution of the phase codes and their respective filter codes, resulting in correlation of the two sequences. In the case where the filters have "matched filter" characteristics (i.e., $a_i' = a_i$) then the sequences are said to be auto-correlated. These statements apply to any pair of equal-length binary sequences. Then if the input sequences utilized are a complementary coded pair, and matched filters are used, the following conditions apply:

$$s_j = c_j + d_j = 0 \qquad \text{for } j \neq 0$$

$$s_o = c_o + d_o = 2n \qquad \text{for } j = 0$$

This is a mathematical description of the final output sequence obtained in the process illustrated above. The final summation terms, s_j, are zero for all values except j = 0, which corresponds to the main pulse output during the eighth time segment.

A second useful property of complementary sequences is that, for each sequence pair, there is an "orthogonal" pair which produces a unique crosscorrelation output. That is,

the output sequences c_j and d_j obtained by correlating a complementary sequence pair with its corresponding orthogonal pair will be described by the relation:

$$s_j = c_j + d_j = 0 \quad \text{for all values of } j.$$

This process would not be useful for pulse radar or data transmission systems, since the processed signal has zero amplitude for all values of time. However, the cross-correlation output is applicable to the RDF system, since it provides a unique correspondence between the angle of arrival and the signal processor output.

The generation of an orthogonal sequence from a given complementary sequence utilizes four of the six basic operations described earlier in this section. The generation algorithm, shown below for an eight-element complementary sequence pair, is given by French [6] in his Master's thesis.

(1) Given a pair of complementary sequences:

A = − − − + − − + −

B = − − − + + + − +

(2) Perform operation #1 (interchange sequences):

A' = − − − + + + − +

B' = − − − + − − + −

(3) Perform operations #2 and #3 (reverse the order of the sequences):

A" = + − + + + − − −

B" = − + − − + − − −

(4) Perform operation #4 (complement elements of A"):

A"' = − + − − − + + +

B"' = − + − − + − − −

The resultant sequence pair A''', B''' is orthogonal to the original sequence pair A,B.

An example of the crosscorrelation process, utilizing orthogonal complementary sequence pairs, is shown in Figures 2-5 and 2-6. The phase coded pulses in Figure 2-5(a) utilize the eight element complementary sequences A and B. The correlation filters, shown in Figure 2-5(b), utilize the orthogonal sequences '''A''' and '''B'''. The outputs from these filters, as shown in Figure 2-6, add to zero for all values of time.

With this introduction to complementary sequence coding and an application to signal processing, the basis is established for its application to the proposed RDF system.

C. APPLICATION OF COMPLEMENTARY SEQUENCE CODING TO AN IDEALIZED RDF SYSTEM

As previously mentioned, the phase coding method of determining the angle of arrival depends upon the time difference between antenna responses in a linear array. This time difference is a unique function of the inter-antenna spacing, and the arrival angle of the wavefront. This is illustrated in Figure 2-7(a), in which a single rectangular RF pulse is assumed incident to a four element array.

The rectangular RF voltage pulses induced at the antenna outputs are separated in the time domain by a shift (Δt) given by the equation:

$$\Delta t = \frac{d}{c} \cos \theta$$

where,

d is the antenna spacing,

c is the propagation velocity of the wavefront,

and θ is the arrival angle of the Poynting vector associated with the wavefront.

Figure 2-7(b) illustrates the relative time spacing of the induced RF voltages, with the output of antenna number one chosen as a reference. For a signal incident upon the antenna system at an arrival angle of zero (along the array baseline) the time difference will be equal to the maximum propagation time between antennas. The time difference will reduce to zero as the arrival angle increases to the point where it is normal to the array baseline.

The type of antenna used in the array is not a critical factor in the RDF operation of the system; any omnidirectional antenna can be used. However, the antennas should be identical in their electrical characteristics (pattern, gain, and impedance), and they should be sufficiently separated to minimize mutual coupling.

The first step in the phase coding technique is the transformation of these time shifted signals into a pair

of complementary sequences that can be interpreted by matched filter decoders. The basic hardware necessary for this process is shown in Figure 2-8. Each antenna feeds a dual output multicoupler device which can provide both a normal phase output and an inverted phase output (relative to the input signal). The outputs from these devices are configured so that their phase states form a complementary sequence pair. For this reason, the array length must correspond to one of the possible complementary sequence code lengths (2, 4, 8, 16, 18, 20, etc.). The four element array was chosen here for simplicity.

After phase coding, the output signals corresponding to each of the complementary sequences are combined in two signal channels and applied to a code interpreter. The code interpreter consists of a pair of correlation filters which, when perfectly matched to these input signals, produce unique autocorrelation (or crosscorrelation) outputs.

To illustrate the phase coding and decoding processes in more detail, the four element array is used with an incident wavefront at an arrival angle of zero degrees. The complementary sequences are identical to those in Figure 2-8, and an idealized rectangular RF pulse is assumed, with a pulse width equal to twice the inter-antenna propagation delay ($2\frac{d}{c}$). To simplify the present example, it is also assumed that the phase reversal and summing operations are performed at the carrier frequency of the incoming pulse signal, and that no other signals are present. The

practical requirements for signal preselection, and inter-signal discrimination circuits are considered in a later section of this chapter.

Based on the above conditions, the signals at the antenna outputs would be as shown in Figure 2-9. The output of antenna number one is again chosen as a reference, and the A & B outputs of each antenna correspond to the channels into which the phase-coded outputs are combined. For convenience, the peak carrier amplitudes of the wave forms are normalized to unity, and upward or downward pointing arrows are used to indicate their respective phase states. An upward pointing arrow indicates a non-inverted output, and a downward pointing arrow indicates an inverted output. When these individual antenna outputs are grouped and then added in an ideal linear (and lossless) summer, the resultant output waveforms are assumed to be as shown in Figure 2-10.

It should be noted that the amplitudes and phases of these idealized waveforms are based on the assumption that the antenna outputs are combined with perfect carrier phase cancellation (180° phase difference) or perfect phase reinforcement (0° phase difference). To assume that perfect phase matching occurs for all possible values of time difference (and, therefore, for all possible values of relative phase difference) is a questionable assumption. That is, perfect 180° or 0° matching between phase coded antenna outputs will occur only for those arrival angles which produce phase differences that are multiples of 2π.

However, the waveforms in Figure 2-10 do provide a simplified illustration of the basic phase coding/decoding technique applied to the RDF system. In addition, it will be shown in the following section that the relative carrier phase differences between individual antenna outputs do not affect the validity of the overall process so long as the signals are combined in a linear device and the principle of superposition can be applied.

Returning then to this simplified example, the coded and summed signals occupying channel A and channel B are each applied to correlation filters which are configured as shown in Figure 2-11. The filters consist of discrete delay devices (or tapped delay lines) and dual output buffer devices with inverted and non-inverted outputs. The dual outputs of the buffer units are configured so that both autocorrelation and crosscorrelation of the coded input signals can be performed. For perfect filter-to-signal matching, the filter time delays must equal the inter-antenna propagation time ($\frac{d}{c}\cos\theta$). For this example, ($\theta = 0°$) perfect matching requires that the time delays equal $\frac{d}{c}$. The autocorrelation and crosscorrelation filter outputs are shown in Figure 2-12 for the perfect matching case. As in Figure 2-9, upward and downward pointing arrows indicate the signal phase states, and the amplitudes correspond to peak values. When the corresponding (autocorrelation and crosscorrelation) outputs of channels A and B are combined in a linear device, the final output waveforms will

be as shown in Figure 2-13. The autocorrelation output, for perfect filter-to-signal matching, is an exact replica of the original RF pulse, with a processing gain equal to twice the code length. The crosscorrelation output is everywhere zero because of the complete phase cancellation of all signal components.

A computer model of this process has been used [1] to determine the effect of correlation filter mismatch upon the autocorrelation and crosscorrelation output waveforms. An important result of that study was the determination that the crosscorrelation output is uniquely zero for the perfectly matched case, and that a very slight amount of mismatch will cause a rapid increase in the crosscorrelation output. It was concluded, then, that the distinct null effect produced by the crosscorrelation process could be utilized to accurately determine the angle of arrival of incident signals.

The computer model utilized in that study employed the idealized signal model that was assumed above, and it also assumed a noise-free signal environment. In order to more fully evaluate the performance of the complementary sequence phase coding process for this paper, the computer model was modified to investigate the effects of background noise and various signal parameters. In addition, the model was utilized to investigate the effects of array length upon the overall process. A description of the model, and the results

and conclusions reached, are presented in Chapter III of this paper.

D. SINUSOIDAL ANALYSIS

The following sinusoidal analysis of the coding/decoding process will show that the basic technique schematically described in the previous section is valid for all arrival angles (between 0° and 90°), and that a 0° or 180° phase relationship between the antenna outputs is not a requirement for proper operation of the system.

First of all, it is assumed that the physical constraints on the system (array size, phase coding sequences, matched filter configuration, and signal parameters) are identical to those specified in the previous example. However, for the purpose of this analysis, the arrival angle may take on any arbitrary value between 0° and 90°.

Therefore, the time differences between antenna responses will be

$$\Delta t \doteq \frac{d}{c} \cos \theta$$

and the relative phase differences will be

$$\Delta\phi = 2\pi \frac{d}{\lambda} \cos \theta = \omega \frac{d}{c} \cos \theta \quad ,$$

where ω is the carrier frequency of the incident signal. Now, if antenna number one is chosen as a time and phase reference, then the signal voltages induced in the four antennas may be represented by the equations:

$$v_1(t) = [u(t)-u(t-2\frac{d}{c})]\cos \omega t = u_1(t)\cos \omega t$$

$$v_2(t) = [u(t-\Delta t)-u(t-2\frac{d}{c}-\Delta t)]\cos(\omega t-\Delta\phi) = u_2(t)\cos(\omega t-\Delta\phi)$$

$$v_3(t) = [u(t-2\Delta t)-u(t-2\frac{d}{c}-2\Delta t)]\cos(\omega t-2\Delta\phi) = u_3(t)\cos(\omega t-2\Delta\phi)$$

$$v_4(t) = [u(t-3\Delta t)-u(t-2\frac{d}{c}-3\Delta t)]\cos(\omega t-3\Delta\phi) = u_4(t)\cos(\omega t-3\Delta\phi)$$

These equations represent the signal waveforms shown in Figure 2-7. The gating functions $u_1(t)$ through $u_4(t)$ define the leading and trailing edges of the signal pulses, which are of duration $2\,\frac{d}{c}$. These signals are then phase coded using complementary sequences and combined in a pair of linear devices so that the resultant outputs are represented by:

Channel A: $v_A = +v_1 + v_2 + v_3 - v_4$

Channel B: $v_B = +v_1 + v_2 - v_3 + v_4$

The composite signals v_A and v_B are then input to auto-correlation and crosscorrelation filters with discrete time delay segments equal to $\frac{d}{c}\cos\theta$. Since the time delays are equivalent to negative phase shifts $(-\Delta\phi)$ equal to $-\omega(\frac{d}{c}\cos\theta)$, then the four correlation outputs can be represented by the series:

Channel A (autocorrelated):

$$V_A = -v_A + v_A\underline{/-\Delta\phi} + v_A\underline{/-2\Delta\phi} + v_A\underline{/-3\Delta\phi}$$

Channel B (autocorrelated):

$$V_B = +v_B - v_B\underline{/-\Delta\phi} + v_B\underline{/-2\Delta\phi} + v_B\underline{/-3\Delta\phi}$$

Channel A (crosscorrelated):

$$V_A' = -v_A - v_A\underline{/-\Delta\phi} + v_A\underline{/-2\Delta\phi} - v_A\underline{/-3\Delta\phi}$$

Channel B (crosscorrelated):

$$V_B' = +v_B + v_B\underline{/-\Delta\phi} + v_B\underline{/-2\Delta\phi} - v_B\underline{/-3\Delta\phi}$$

where $v_X\underline{/-n\Delta\phi}$ represents the composite voltage (v_A or v_B) phase shifted by an angle $-n\Delta\phi$. For example, expanding $v_A\underline{/-n\Delta\phi}$ produces the following equation:

$$v_A\underline{/-n\Delta\phi} = v_1\underline{/-n\Delta\phi} + v_2\underline{/-n\Delta\phi} + v_3\underline{/-n\Delta\phi} - v_4\underline{/-n\Delta\phi}$$

A general expression for the series terms is

$$v_k\underline{/-n\Delta\phi} = \left\{u_k(t)\ \cos[\omega t-(k-1)\Delta\phi]\right\}\ \underline{/-n\Delta\phi} \qquad \begin{matrix}(k=1,2,3,4)\\(n=0,1,2,3)\end{matrix}$$

where $u_k(t) = u[t-(k-1)\Delta t] - u[t-2\frac{d}{c}-(k-1)\Delta t]$.

Since the phase delay $\underline{/-n\Delta\phi}$ is equivalent to the time delay $n\Delta t$, the gating function $u_k(t)$ is delayed by $n\Delta t$, and becomes:

$$u_k(t+n\Delta t) = u[t-(k+n-1)\Delta t] - u[t-2\frac{d}{c}-(k+n-1)\Delta t] = u_{k+n}(t).$$

Therefore, the general series term may be expressed as:

$$v_k\underline{/-n\Delta\phi} = u_{k+n}(t)\cos[\omega t-(k+n-1)\Delta\phi]$$

Now, applying these terms to the autocorrelation and cross-correlation series:

$$\begin{array}{llllll}
 & (k=1) & (k=2) & (k=3) & (k=4) & \\
V_A = & -u_1\cos(\omega t) & -u_2\cos(\omega t-\Delta\phi) & -u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & (n=0)\\
 & +u_2\cos(\omega t-\Delta\phi) & +u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & -u_5\cos(\omega t-4\Delta\phi) & (n=1)\\
 & +u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & +u_5\cos(\omega t-4\Delta\phi) & -u_6\cos(\omega t-5\Delta\phi) & (n=2)\\
 & -u_4\cos(\omega t-3\Delta\phi) & +u_5\cos(\omega t-4\Delta\phi) & +u_6\cos(\omega t-5\Delta\phi) & -u_7\cos(\omega t-6\Delta\phi) & (n=3)
\end{array}$$

$$= -u_1\cos(\omega t)+u_3\cos(\omega t-2\Delta\phi)+4u_4\cos(\omega t-3\Delta\phi)+u_5\cos(\omega t-4\Delta\phi)-u_7\cos(\omega t-6\Delta\phi).$$

$$\begin{array}{llllll}
 & (k=1) & (k=2) & (k=3) & (k=4) & \\
V_B = & +u_1\cos(\omega t) & +u_2\cos(\omega t-\Delta\phi) & -u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & (n=0)\\
 & -u_2\cos(\omega t-\Delta\phi) & -u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & -u_5\cos(\omega t-4\Delta\phi) & (n=1)\\
 & +u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & -u_5\cos(\omega t-4\Delta\phi) & +u_6\cos(\omega t-5\Delta\phi) & (n=2)\\
 & +u_4\cos(\omega t-3\Delta\phi) & +u_5\cos(\omega t-4\Delta\phi) & -u_6\cos(\omega t-5\Delta\phi) & +u_7\cos(\omega t-6\Delta\phi) & (n=3)
\end{array}$$

$$= u_1\cos(\omega t)-u_3\cos(\omega t-2\Delta\phi)+4u_4\cos(\omega t-3\Delta\phi)-u_5\cos(\omega t-4\Delta\phi)+u_7\cos(\omega t-6\Delta\phi).$$

$$\begin{array}{llllll}
 & (k=1) & (k=2) & (k=3) & (k=4) & \\
V_A' = & -u_1\cos(\omega t) & -u_2\cos(\omega t-\Delta\phi) & -u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & (n=0)\\
 & -u_2\cos(\omega t-\Delta\phi) & -u_3\cos(\omega t-2\Delta\phi) & -u_4\cos(\omega t-3\Delta\phi) & +u_5\cos(\omega t-4\Delta\phi) & (n=1)\\
 & +u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & +u_5\cos(\omega t-4\Delta\phi) & -u_6\cos(\omega t-5\Delta\phi) & (n=2)\\
 & -u_4\cos(\omega t-3\Delta\phi) & -u_5\cos(\omega t-4\Delta\phi) & -u_6\cos(\omega t-5\Delta\phi) & +u_7\cos(\omega t-6\Delta\phi) & (n=3)
\end{array}$$

$$= -u_1\cos(\omega t)-2u_2\cos(\omega t-\Delta\phi)-u_3\cos(\omega t-2\Delta\phi)+u_5\cos(\omega t-4\Delta\phi)-2u_6\cos(\omega t-5\Delta\phi)$$
$$+u_7\cos(\omega t-6\Delta\phi).$$

$$
\begin{array}{lllll}
 & (k=1) & (k=2) & (k=3) & (k=4) \\
V_B' = & +u_1\cos(\omega t) & +u_2\cos(\omega t-\Delta\phi) & -u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) \quad (n=0) \\
 & +u_2\cos(\omega t-\Delta\phi) & +u_3\cos(\omega t-2\Delta\phi) & -u_4\cos(\omega t-3\Delta\phi) & +u_5\cos(\omega t-4\Delta\phi) \quad (n=1) \\
 & +u_3\cos(\omega t-2\Delta\phi) & +u_4\cos(\omega t-3\Delta\phi) & -u_5\cos(\omega t-4\Delta\phi) & +u_6\cos(\omega t-5\Delta\phi) \quad (n=2) \\
 & -u_4\cos(\omega t-3\Delta\phi) & -u_5\cos(\omega t-4\Delta\phi) & +u_6\cos(\omega t-5\Delta\phi) & -u_7\cos(\omega t-2\Delta\phi) \quad (n=3)
\end{array}
$$

$$
= u_1\cos(\omega t) + 2u_2\cos(\omega t-\Delta\phi) + u_3\cos(\omega t-2\Delta\phi) - u_5\cos(\omega t-4\Delta\phi) + 2u_6\cos(\omega t-5\Delta\phi) - u_7\cos(\omega t-6\Delta\phi).
$$

Finally, when each channel's autocorrelation and cross-correlation terms are combined, the resulting outputs will be:

$$
V_o = V_A + V_B = 8u_4\cos(\omega t-3\Delta\phi)
$$

$$
= 8\left\{u[t-3\Delta t] - u[t - 2\,\frac{d}{c} - 3\Delta t]\right\}\cos(\omega t-3\Delta\phi)
$$

$$
V_o = V_A' + V_B' = 0
$$

These equations represent the final output waveforms shown schematically in Figure 2-13. The autocorrelation output is an exact replica of the incoming pulse signal, and is amplified by a factor of twice the code length. The gating function $u_4(t)$ corresponds to the result in Figure 2-13 in that the leading edge of the output pulse is delayed $3\Delta t$ time units from the initial intercept of the incident pulse. Also, as in the graphical example, the final crosscorrelation output is everywhere zero. Therefore, it is concluded that the phase coding process is valid for all

arrival angles between 0° and 90°, that it provides a signal processing gain (in the autocorrelation mode) of twice the chosen code length, and that it provides a crosscorrelation output that is everywhere zero.

E. ENGINEERING CONSIDERATIONS

In order to apply the basic phase coding technique described above to a physically realizable RDF system, certain practical problem areas must be considered. During the investigation and evaluation of this technique, it became obvious that several factors have a direct effect upon the operation of the proposed system. Specifically, those factors that were considered most important are: (1) possible sources of error resulting from angular ambiguities, (2) the requirement for signal selection and discrimination, and (3) the effect of different modulation types on the correlation processes. The constraints and physical requirements which these factors place upon the design of such a system will, in large part, determine the feasibility of utilizing this type of RDF system over some other type.

The first two factors (angular ambiguity and signal preselection) are discussed and evaluated in this section. The actual design requirements that must be imposed to satisfy these two areas are straightforward, and are considered to be technically feasible. However, the overall practicality or economic feasibility of these requirements

will depend, in part, upon the particular application of the system, and are not addressed in this paper. Modulation effects are best described and evaluated by the computer model used to simulate the phase coding and decoding portion of the system. This discussion is presented in the following chapter.

1. Angular Ambiguities

Until now, the determination of arrival angle has been assumed to be a two-dimensional problem, in which one single-variable equation is evaluated to yield a single valued unambiguous solution. However, in most environmental situations, the direction of arrival of an incident wavefront will generally be composed of both azimuthal and elevation angle of arrival components. The additional consideration of elevation angle introduces a third spatial dimension to the previous two dimensional problem. That is, the elevation angle introduces a second variable to the time delay equation, so that $\Delta t = \frac{d}{c} \cos \theta \cos \alpha$, where α is the elevation angle of arrival. This additional variable precludes the unambiguous determination of azimuthal angle from the single time difference equation.

Furthermore, quadrantal ambiguities exist because of the periodic nature of the cosine function. For example, azimuth arrival angles of 45°, 135°, 225°, and 315° would produce equal magnitude time differences between adjacent antenna responses. Fortunately, some differentiation between these responses is possible by knowing the algebraic

sign of the time differences. This differentiation can be accomplished by a system using the phase coding technique because the chronological order in which the individual antennas respond to the incident wavefront will uniquely determine the composition of the phase coded sequences. For example, two incident wavefronts whose azimuthal arrival angles differ by 180° will create distinctively different phase coded sequences. Correlation filter pairs which are matched to either one of these sequences will not be matched to the other.

This can be demonstrated using the example illustrated in Figures 2-8 and 2-11, with the rectangular pulse approaching from 180° instead of 0°. Assuming the same phase coding of the antenna signals, the combined antenna outputs for the 180° AOA signal will be as shown in Figure 2-14. When those signals are input to the autocorrelation and crosscorrelation filters (which remain matched for 0° AOA signals) the output waveforms will be as shown in Figure 2-15. Combining the respective autocorrelation and crosscorrelation outputs from the channel A and channel B filters, the final output waveforms will be as shown in Figure 2-16. These outputs indicate a severe mismatch between the coded signals and the correlation filters. Therefore, there is no sense ambiguity in the response of the system. In an actual RDF system utilizing this technique it would be possible to establish a bank of correlation filters which would respond (uniquely) to a full 180° range of azimuthal arrival angles.

However, there are still ambiguities created by azimuthal angles which produce equal (sign and magnitude) time differences. These ambiguities, and the ambiguities created by elevation angles of arrival must be resolved by an interferometer technique. A pair of linear arrays, situated in the ground plane and separated by a known fixed angle, will yield distinct, simultaneous time difference equations related to their respective baselines. A brief analysis of the two-array system, which is represented in Figure 2-17, will show that all spatial ambiguities can be resolved.

The time difference equations which describe the (orthogonal) arrays in Figure 2-17 are:

$$(\Delta t)_1 = \frac{d}{c} \cos \theta_1 \cos \alpha$$

$$(\Delta t)_2 = \frac{d}{c} \cos \theta_2 \cos \alpha$$

where the subscripted time shifts $(\Delta t)_1$ and $(\Delta t)_2$ are introduced by the corresponding linear arrays, and α represents the fixed elevation angle, which may take on any values between 0° and 90°. The azimuth angles θ_1 and θ_2 are related by the orientation of the two linear arrays. For example, in the case of the orthogonal arrays in Figure 2-17;

$$\theta_2 = \theta_1 + 270°.$$

These three equations will yield unambiguous solutions for the azimuth and elevation angles if the amplitude and sign of $(\Delta t)_1$ and $(\Delta t)_2$ can be determined. The uniqueness of these equations can be shown using the four incident wavefronts represented in Figure 2-17 by the vectors I, II, III, and IV. An arbitrary elevation angle (α) is assumed for all four wavefronts, and their azimuth angles are chosen as 30°, 150°, 210°, and 330° relative to the axis of antenna number one. The time difference equations corresponding to these four wavefronts are as follows:

Case I. $(\theta_1 = 30°, \theta_2 = 300°)$

$$(\Delta t)_1 = \frac{d}{c} \cos \theta_1 \cos \alpha = +0.87 \frac{d}{c} \cos \alpha$$

$$(\Delta t)_2 = \frac{d}{c} \cos \theta_2 \cos \alpha = +0.50 \frac{d}{c} \cos \alpha$$

Case II. $(\theta_1 = 150°, \theta_2 = 60°)$

$$(\Delta t)_1 = \frac{d}{c} \cos \theta_1 \cos \alpha = -0.87 \frac{d}{c} \cos \alpha$$

$$(\Delta t)_2 = \frac{d}{c} \cos \theta_2 \cos \alpha = +0.50 \frac{d}{c} \cos \alpha$$

Case III: $(\theta_1 = 210°, \theta_2 = 120°)$

$$(\Delta t)_1 = \frac{d}{c} \cos \theta_1 \cos \alpha = -0.87 \frac{d}{c} \cos \alpha$$

$$(\Delta t)_2 = \frac{d}{c} \cos \theta_2 \cos \alpha = -0.50 \frac{d}{c} \cos \alpha$$

Case IV: $(\theta_1 = 330°, \theta_2 = 240°)$

$$(\Delta t)_1 = \frac{d}{c} \cos \theta_1 \cos \alpha = +0.87 \frac{d}{c} \cos \alpha$$

$$(\Delta t)_2 = \frac{d}{c} \cos \theta_2 \cos \alpha = -0.50 \frac{d}{c} \cos \alpha$$

Although the corresponding time shifts introduced by each of the wavefronts have equal magnitudes, their algebraic signs are unique pair combinations. Therefore, differentiation between these (otherwise ambiguous) wavefronts can be performed once the algebraic sign of each time shift is known. That is, each pair of time difference values corresponds uniquely to a given azimuth angle and elevation angle combination.

It has already been shown that the "sense" and therefore, the algebraic sign of the time shifts can be determined by the matched filter correlation process. Therefore, the two array system suggested here is theoretically capable of determining the parameters necessary to locate the incident wavefront in three dimensional space. However, the design of a specific mechanism to accomplish this is not attempted in this paper. The design, modeling, and evaluation of a physically realizable, three-dimensional RDF system utilizing phase coded linear arrays is considered a sufficient task for additional thesis work.

2. Signal Pre-Selection

As in any radio frequency receiving systems, the radio direction finder must be capable of electronically sorting through a range of operating frequencies in order to select particular signals of interest. In most RDF systems currently in use, pre-selection and identification of signals is accomplished, either by a human operator or automatically, before the line of bearing is obtained for that signal. The idealized RDF system presented in this paper has thus far been limited to operation in a single-signal environment. However, in order to utilize the phase coding technique in a realizable system capable of operating in a dense RF environment, signal preselection and amplification functions must be incorporated into the signal processing portion of the system.

The pre-selection method which is used almost universally in communications receivers is the superheterodyne technique. That is, the desired signal frequency is converted by means of a non-linear mixing process to some intermediate frequency (IF) at which narrowband high gain amplifiers may be used. In some cases the frequency converter circuit is preceded by one or more tuned RF amplifiers to provide coarse frequency selection, and to improve the receiver noise figure. However, the overall receiver selectivity is determined almost entirely by the IF circuits employed.

In order to apply the superheterodyne principle to the proposed RDF system, a decision must be made concerning where the frequency conversion should occur in the signal coding and decoding process. This decision will influence the type of circuits and components that would be required in the implementation of such a system, and in some cases may determine whether or not the system is operationally feasible. Basically, the conversion must occur at a stage of signal processing where it will not affect the RDF operation of the system. There are several locations in the signal path where frequency conversion and envelope detection might take place.

The first possibility involves the phase coding and correlation of all input signals at the carrier frequency, with signal selection, conversion, and envelope detection at the output of the matched filters. The detected outputs of the matched filters would then be scanned for a unique crosscorrelation output. This approach has been utilized in the discussion and analysis of the phase coded system presented thus far in this paper. The sinusoidal analysis in the previous section has shown that phase coding and matched filter correlation of signals can be performed at the carrier frequency for all arrival angles.

The primary disadvantage of this approach is that the signal processing is performed at very low signal levels. Accurate 180° phase inversion and amplitude balance of antenna outputs over a wide range of frequencies would

seem to be a very difficult requirement, considering that signal levels are typically in the microvolt range. These signal levels would be further reduced by the correlation filters, since most time delay devices, such as coaxial cable, electromechanical delay lines, and acoustic surface wave devices, are passive, and introduce signal losses.

In addition, it is not clear what effect many different signals at various frequencies and arrival angles, will have on the phase coding and correlation processes. The possibility of signal interactions due to circuit non-linearities, which are always present in active devices, indicates that ambiguous results or spurious responses might be introduced by this wide-band approach.

This approach does have the advantage however, that only one receiver is required. That is, the various antenna signals are eventually combined into a single channel which could then be input to a standard communications receiver. In this case the scanning process could be performed by electronically sampling the matched filter outputs and feeding the samples to the single receiver.

A second possible configuration of the signal processor involves the frequency conversion of signals at each of the antenna outputs. With this procedure, the low level RF signals are converted to IF signals before the phase coding/decoding process occurs. In this case, the single-signal situation may again be assumed, with the coding/decoding being performed at the intermediate frequency instead of the carrier frequency.

This approach would seem to offer several advantages over the previous wide-open front end approach. First of all, signal preselection before phase coding and correlation processing will eliminate the possibility of interaction between signals. Second, the noise figure of the system will be established by the individual array receivers, so the losses introduced by the correlation process will have negligible effect upon the overall sensitivity of the system. Finally, since the phase coding and correlation process will be performed on high level, fixed frequency signals, more efficient narrowband circuits can be employed.

The primary disadvantage of this system is that it requires the use of multiple receivers, which must be gain and phase matched over their total frequency range. However, it will be shown in the next chapter, that the use of large arrays is not necessary for good RDF resolution capability. Therefore, the use of short (2 or 4 element) array with this approach might result in a satisfactory trade-off between system performance and complexity. This approach, utilizing individual receivers at each antenna, and phase coding and correlation of signals at a fixed intermediate frequency, is employed in the system model evaluated (by computer simulation) in the following chapter.

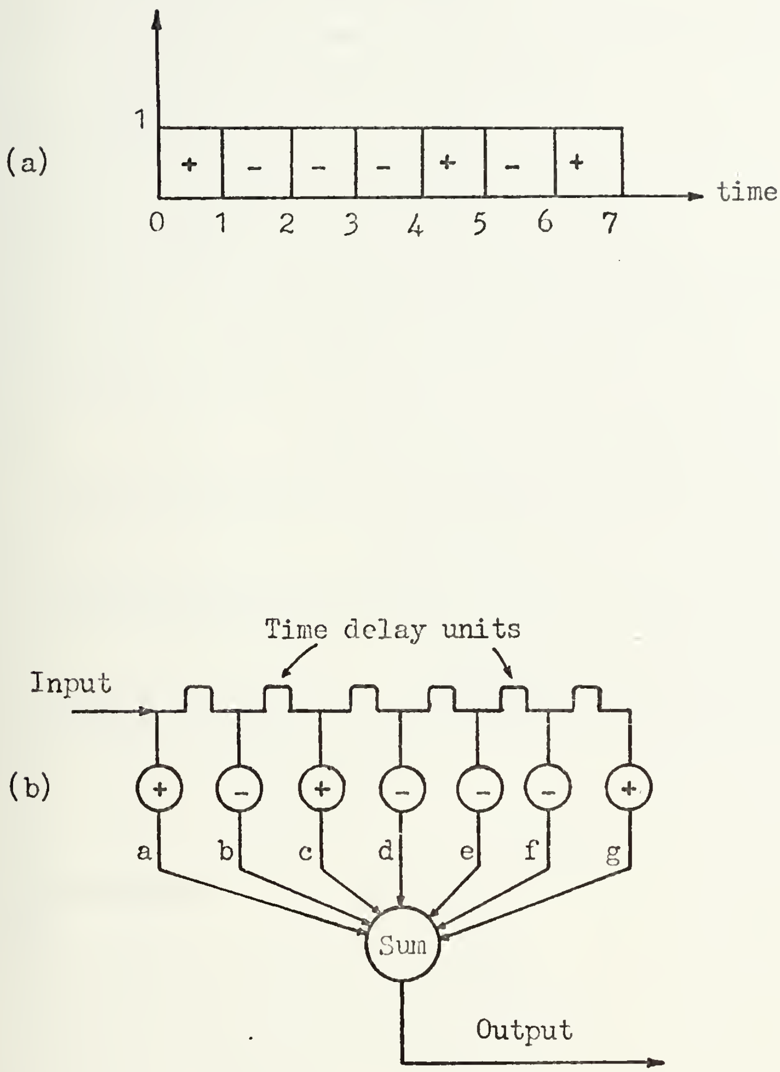


Figure 2-1, Phase Coded Input Pulse and Matched Filter

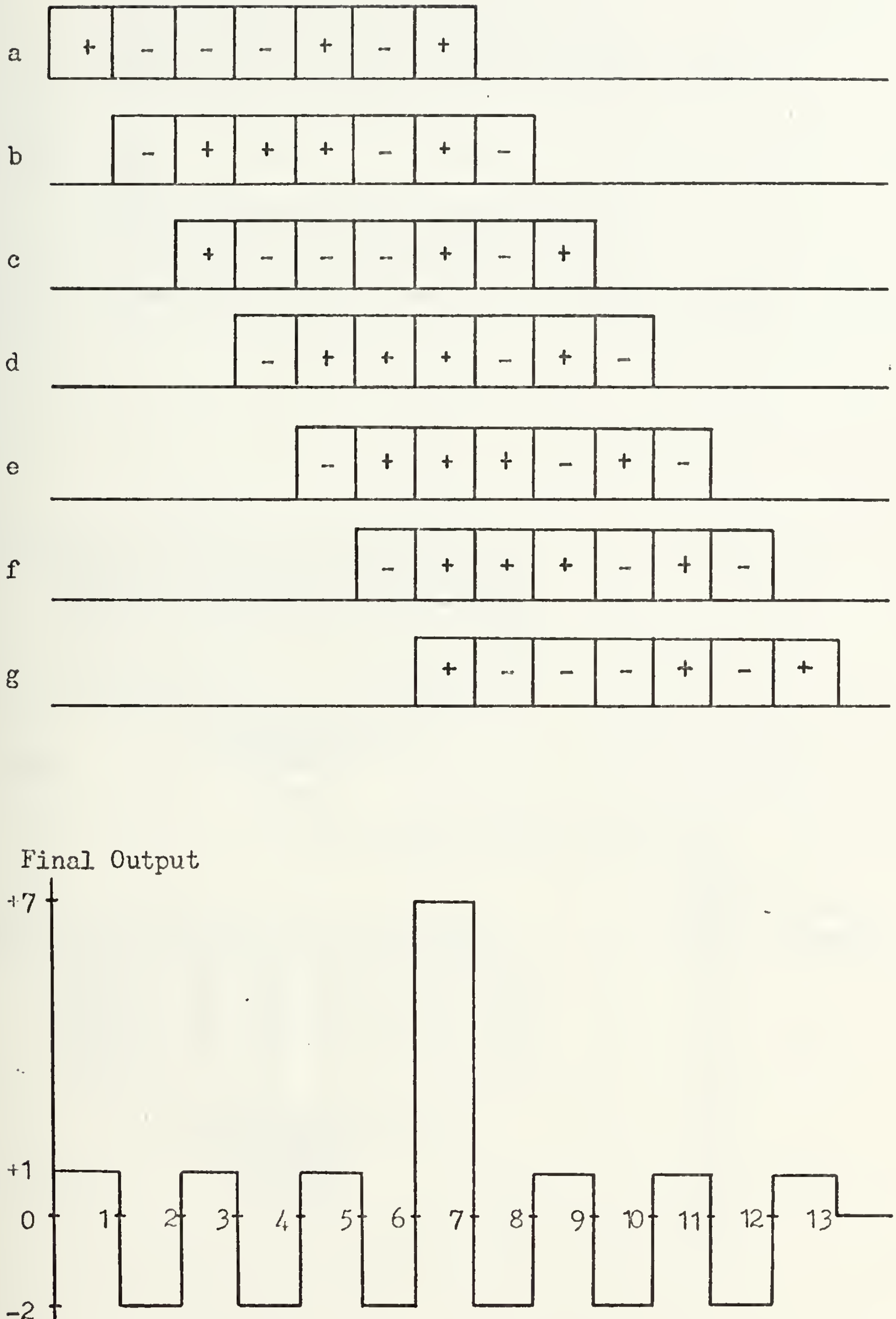


Figure 2-2, Matched Filter Output Waveforms

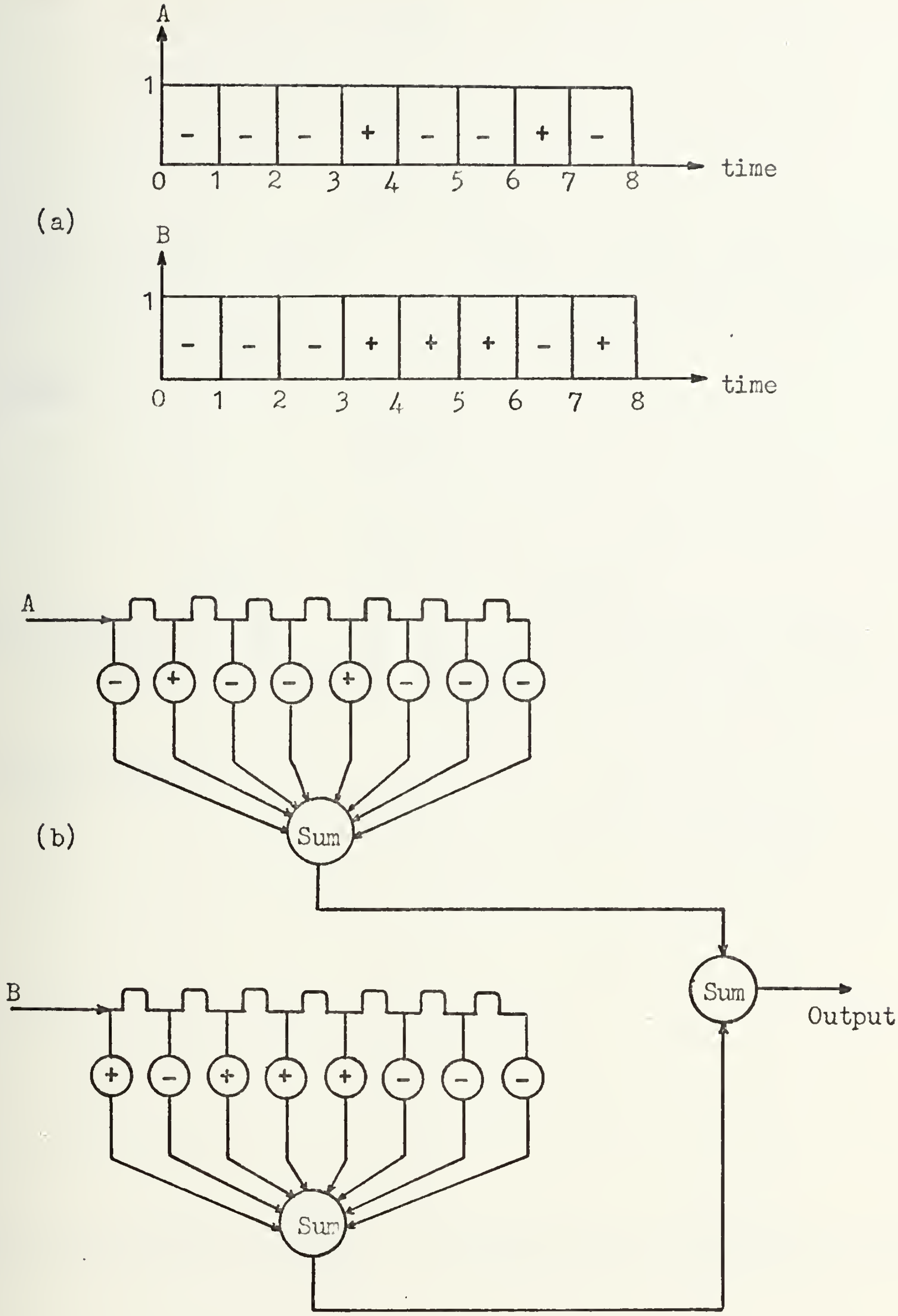


Figure 2-3, Complementary Sequence Phase Coded Pulses and Matched Filters

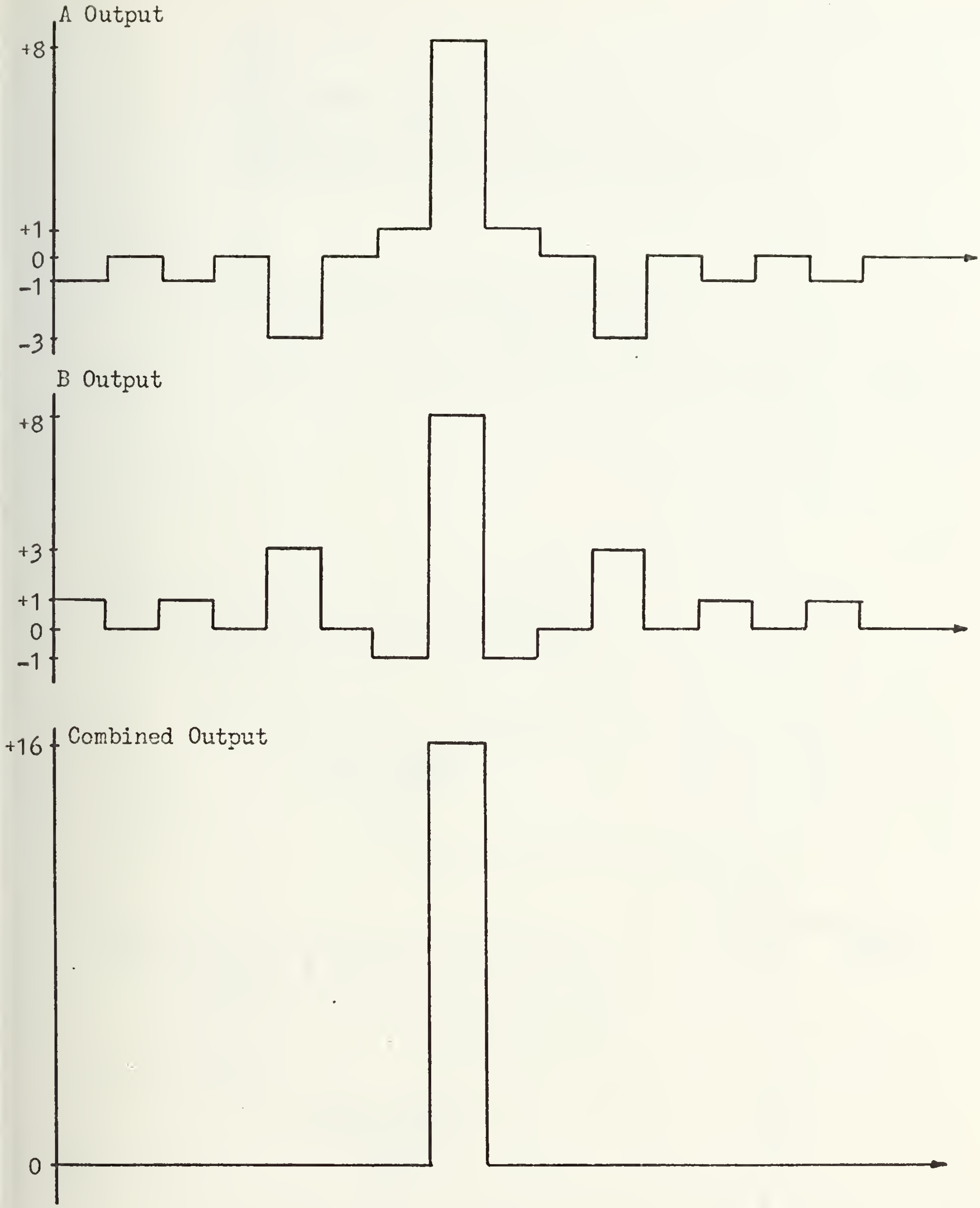


Figure 2-4, Matched Filter Output Waveforms

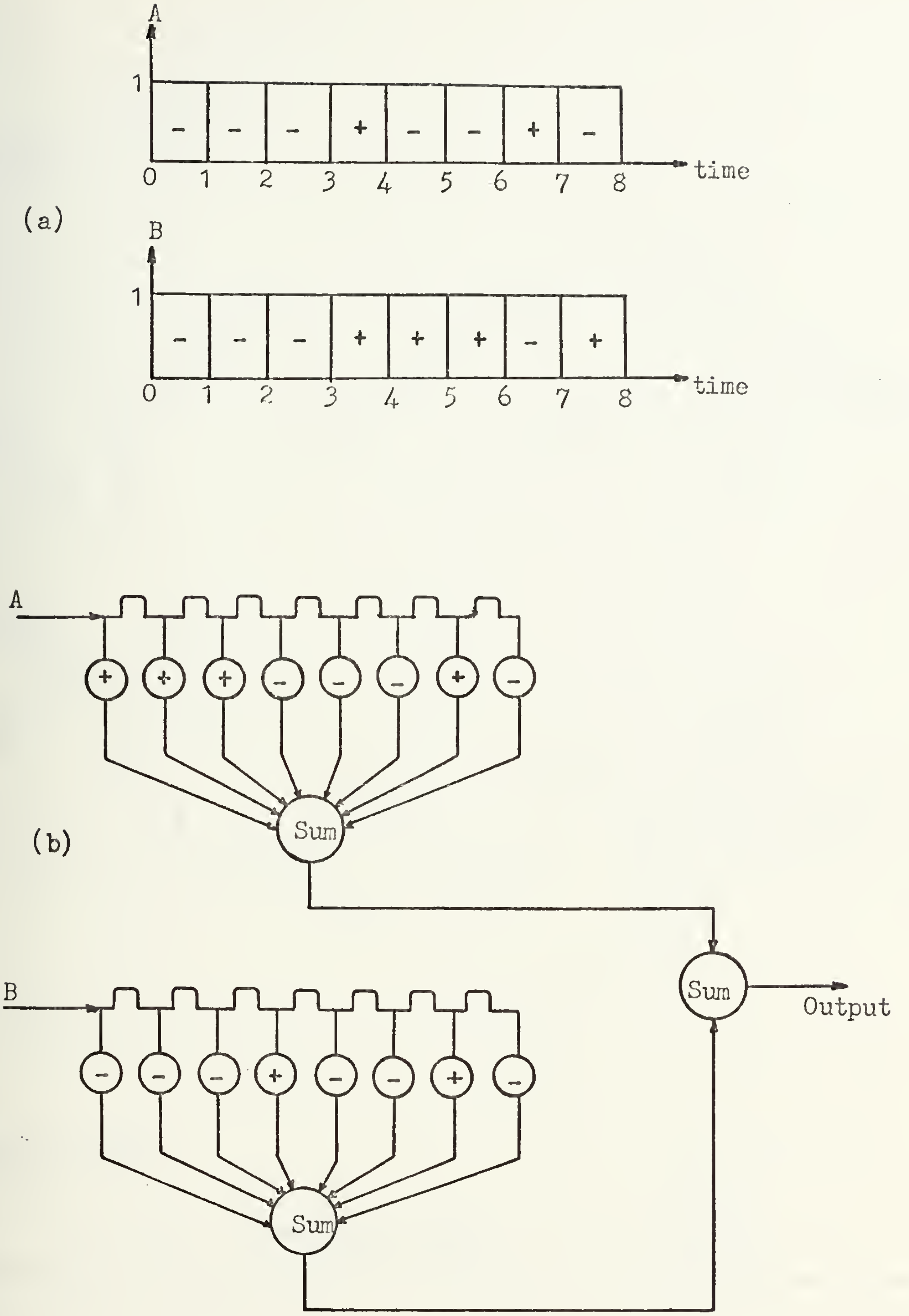


Figure 2-5, Complementary Sequence Phase Coded Pulses and Matched Crosscorrelation Filters

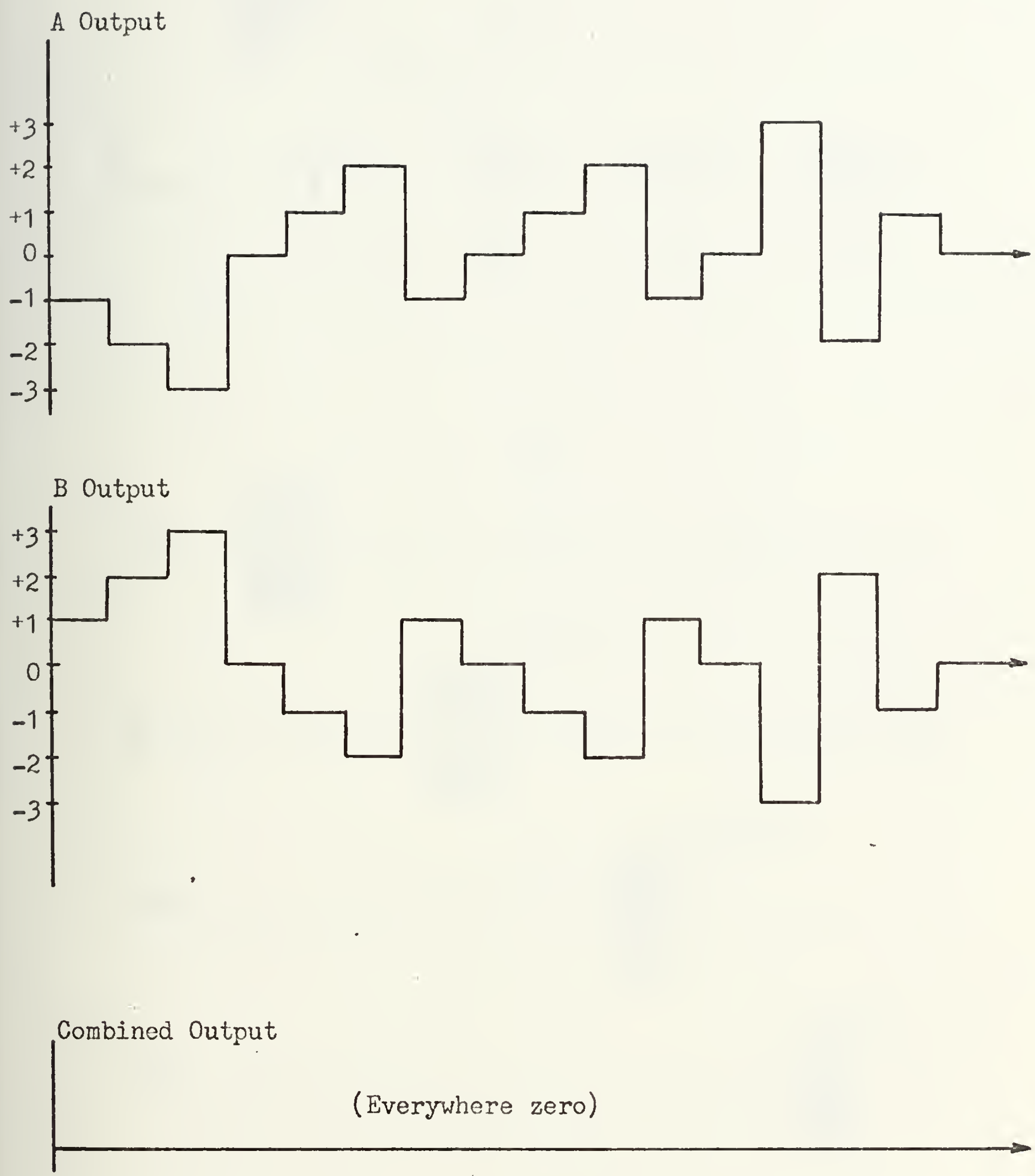


Figure 2-6, Crosscorrelation Filter Output Waveforms

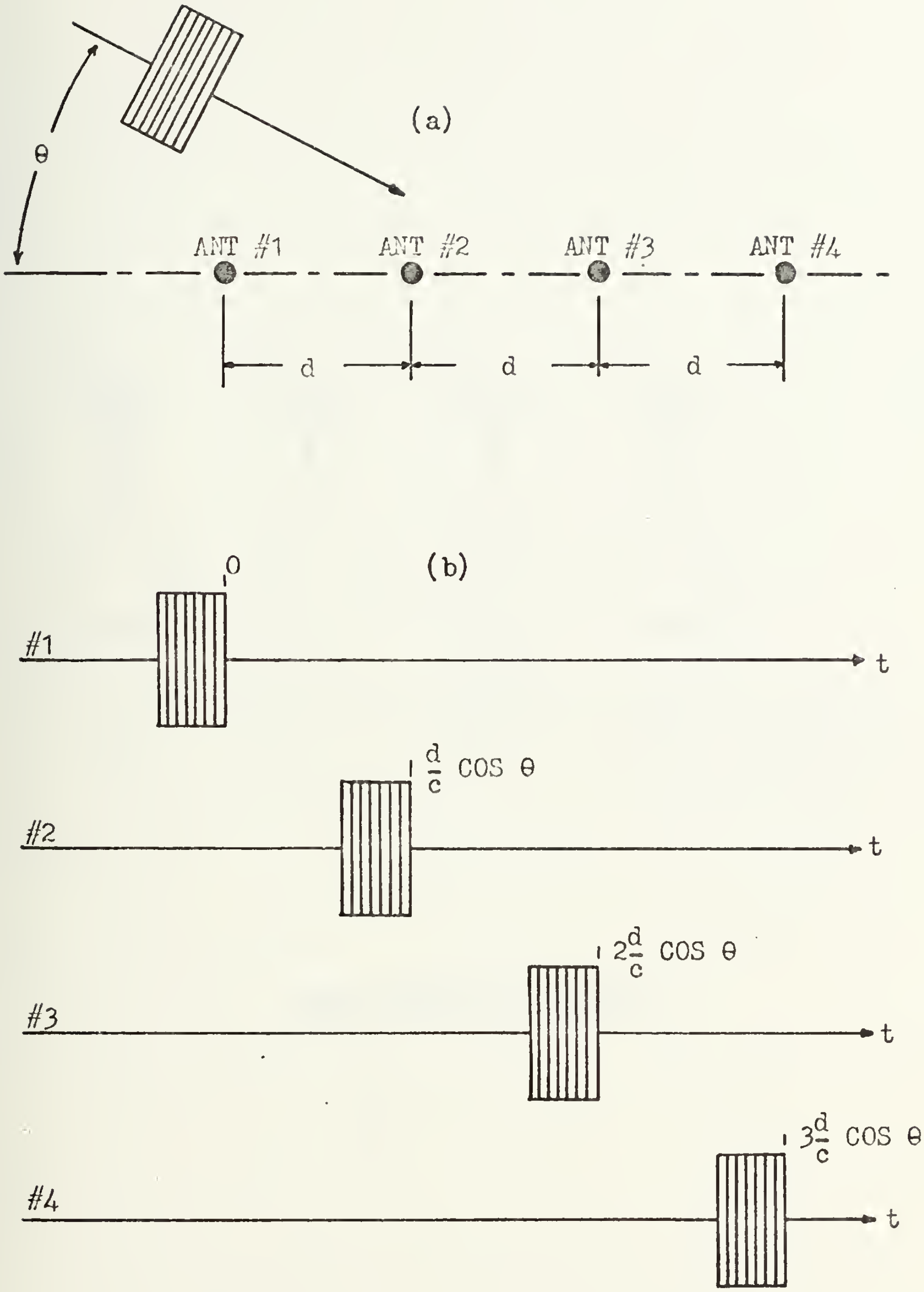
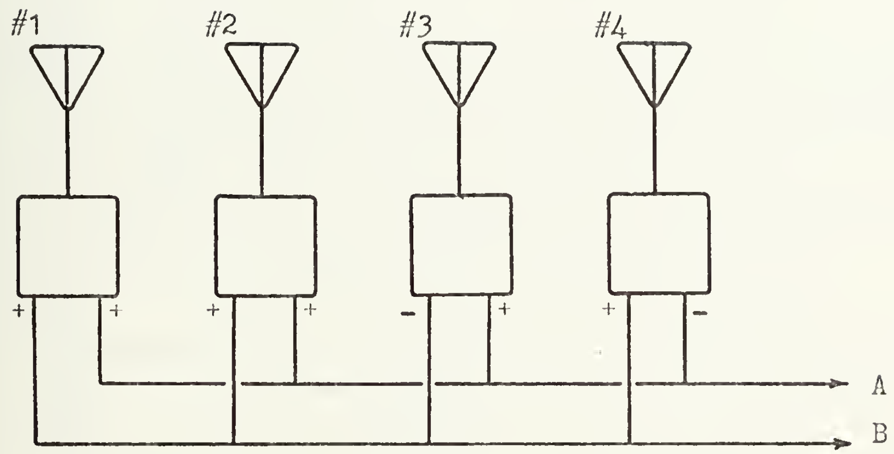


Figure 2-7, Four Element Antenna Array and Induced Signals



COMPLEMENTARY SEQUENCES

A: + + + -

B: + + - +

Figure 2-8, Antenna Output Phase Coding Scheme

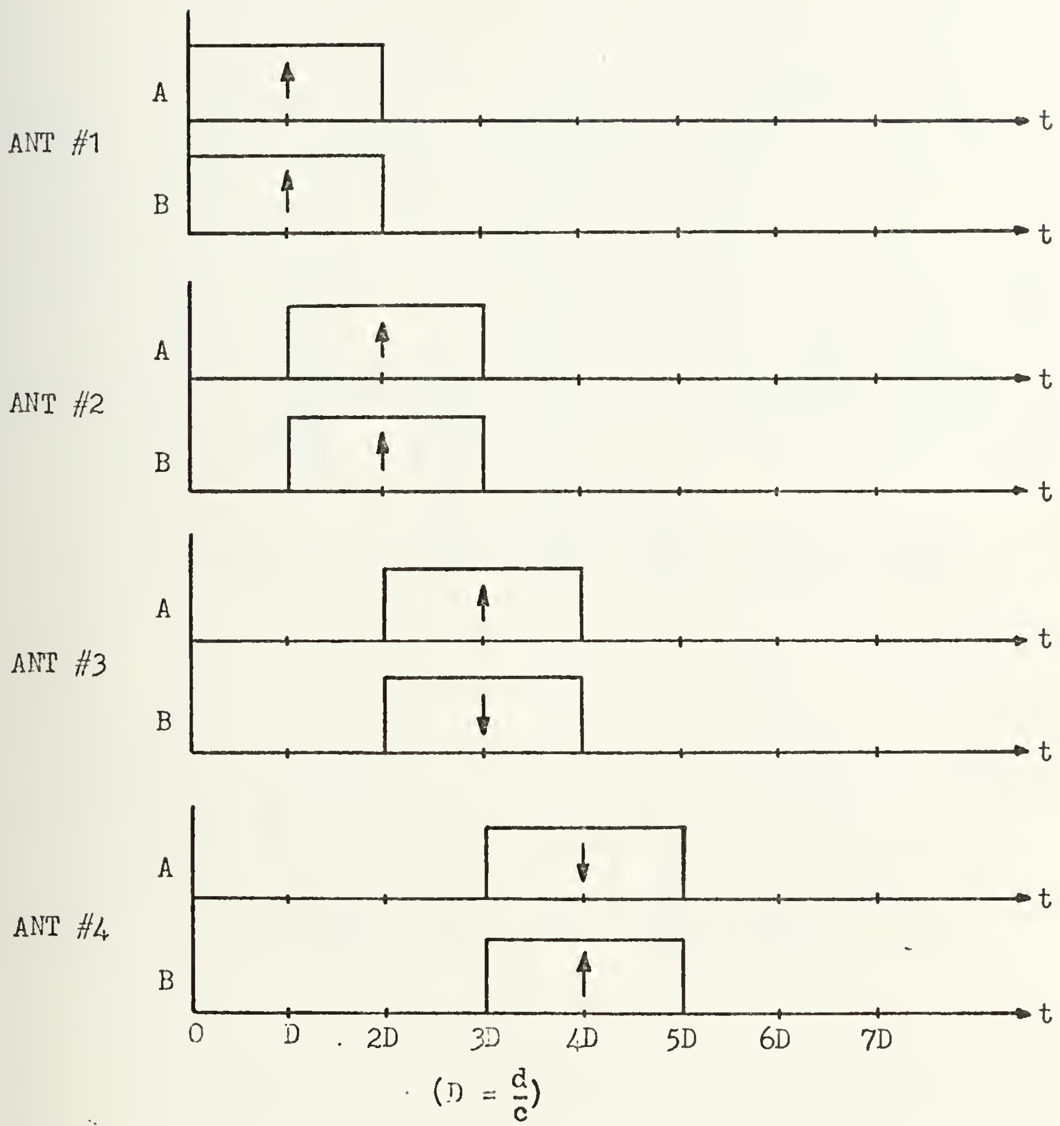


Figure 2-9, Antenna Output Waveforms

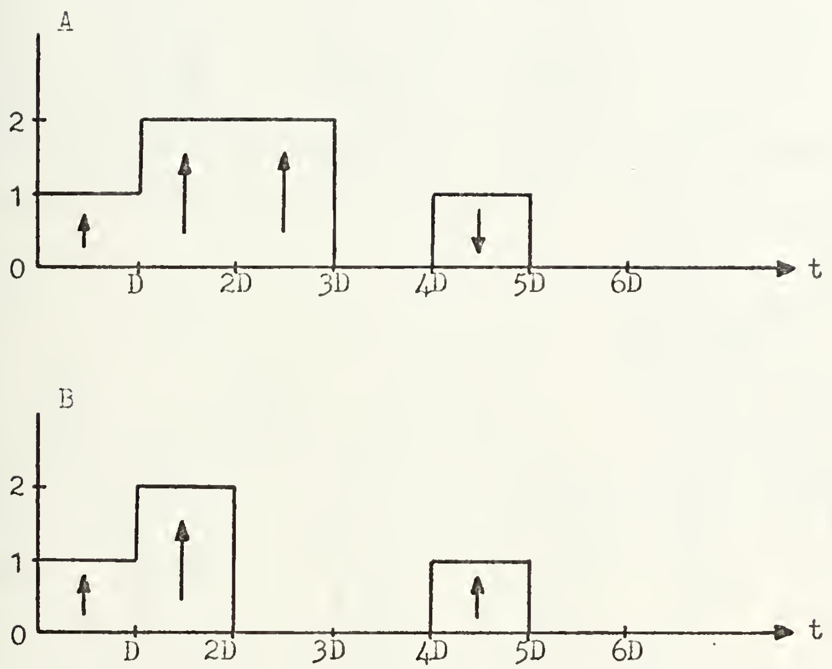
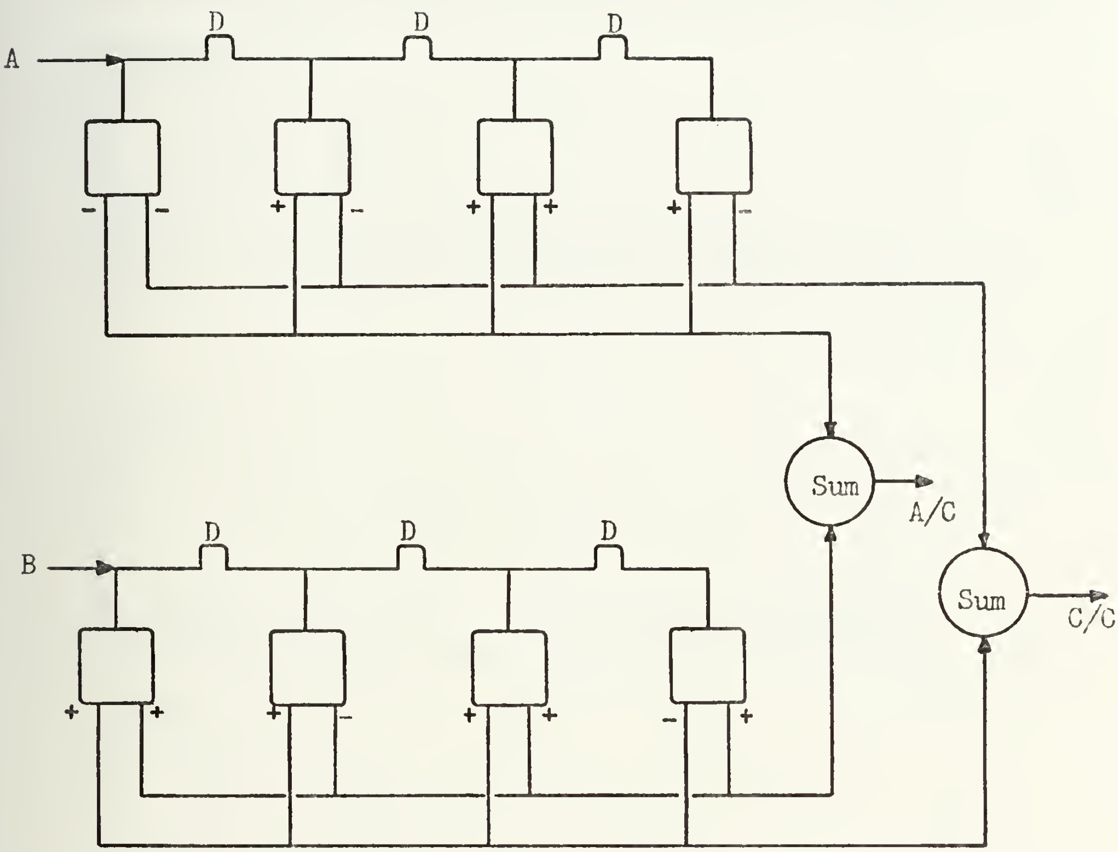


Figure 2-10, Combined Output Waveforms



A/C - Autocorrelation Output

C/C - Crosscorrelation Output

Figure 2-11, Four Element Correlation Filters

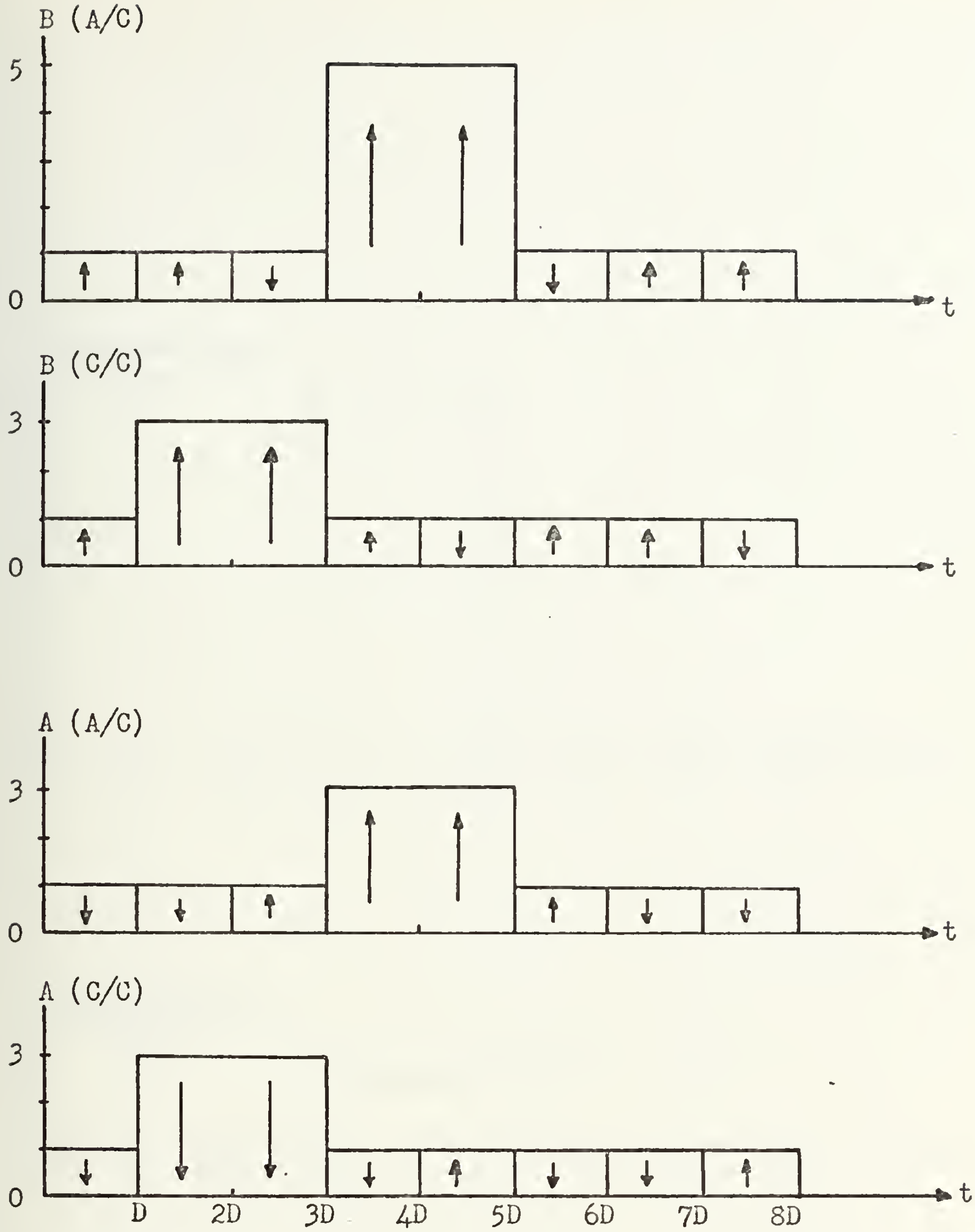


Figure 2-12, Correlation Filter Output Waveforms

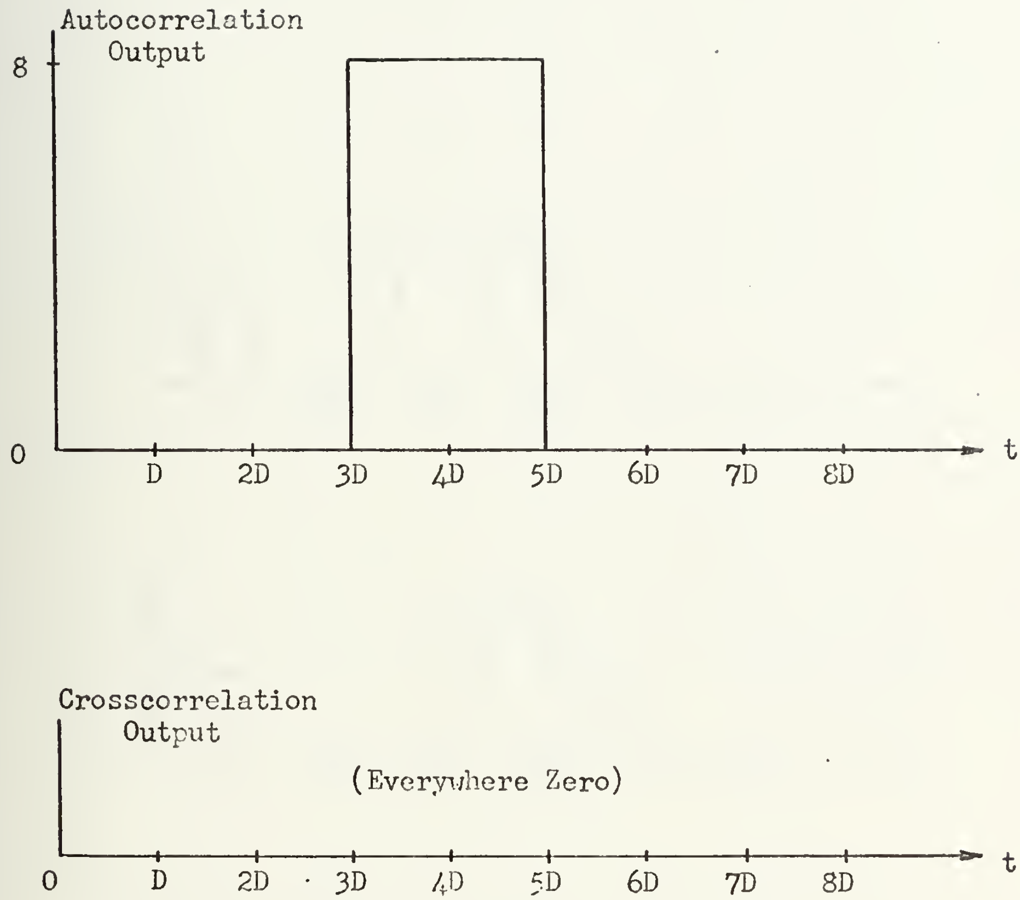


Figure 2-13, Final Autocorrelation and Crosscorrelation Outputs

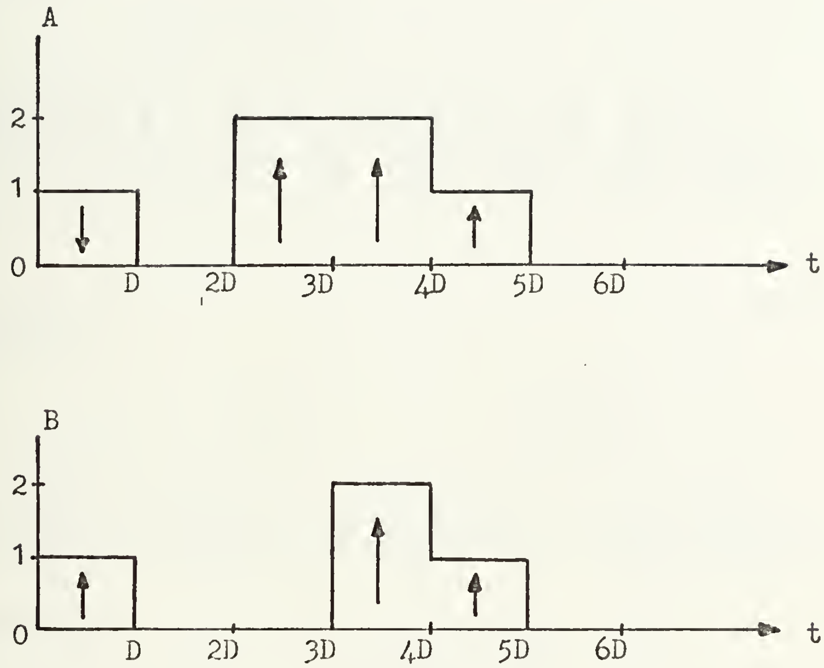


Figure 2-14, Combined Antenna Outputs for 180° Arrival Angle

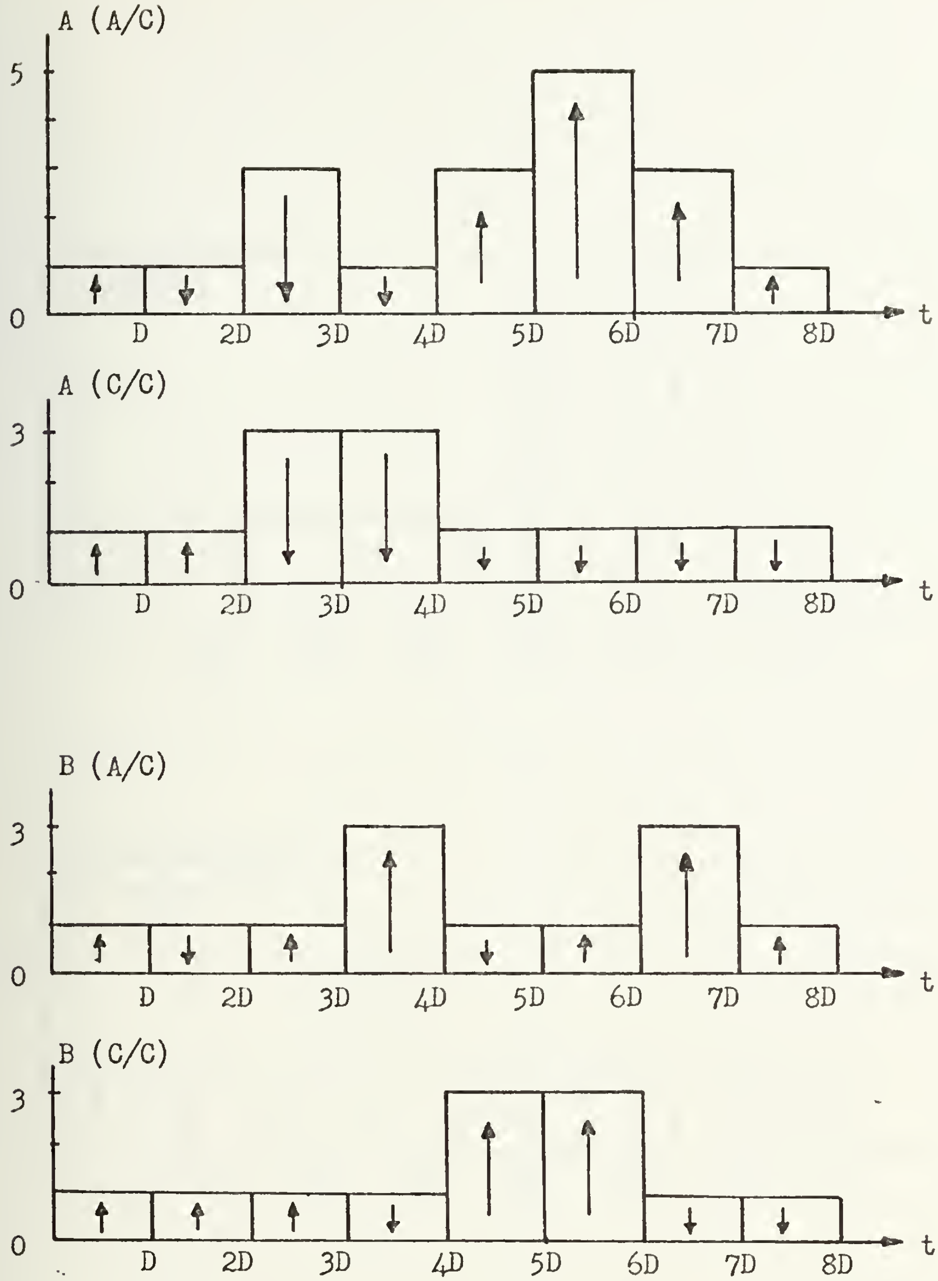


Figure 2-15, Correlation Filter Outputs for 180° Mismatch

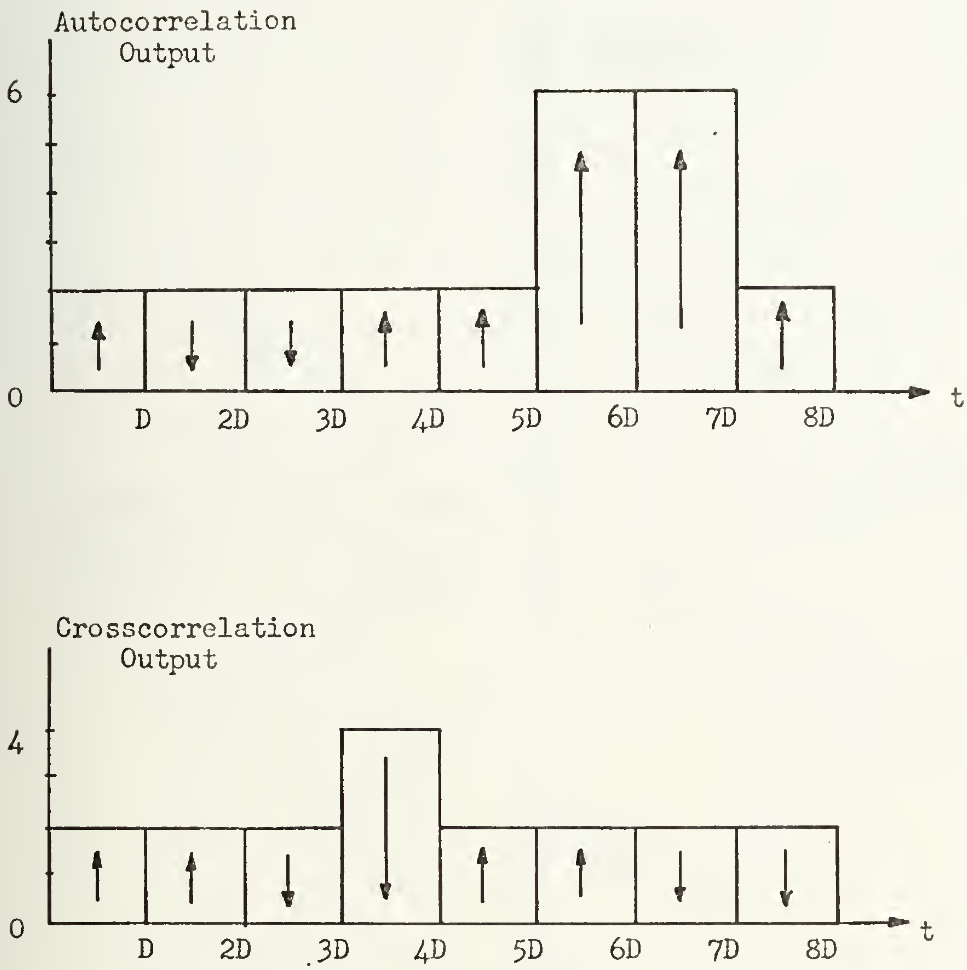


Figure 2-16, Final Autocorrelation and Crosscorrelation Outputs
For a 180° Mismatch

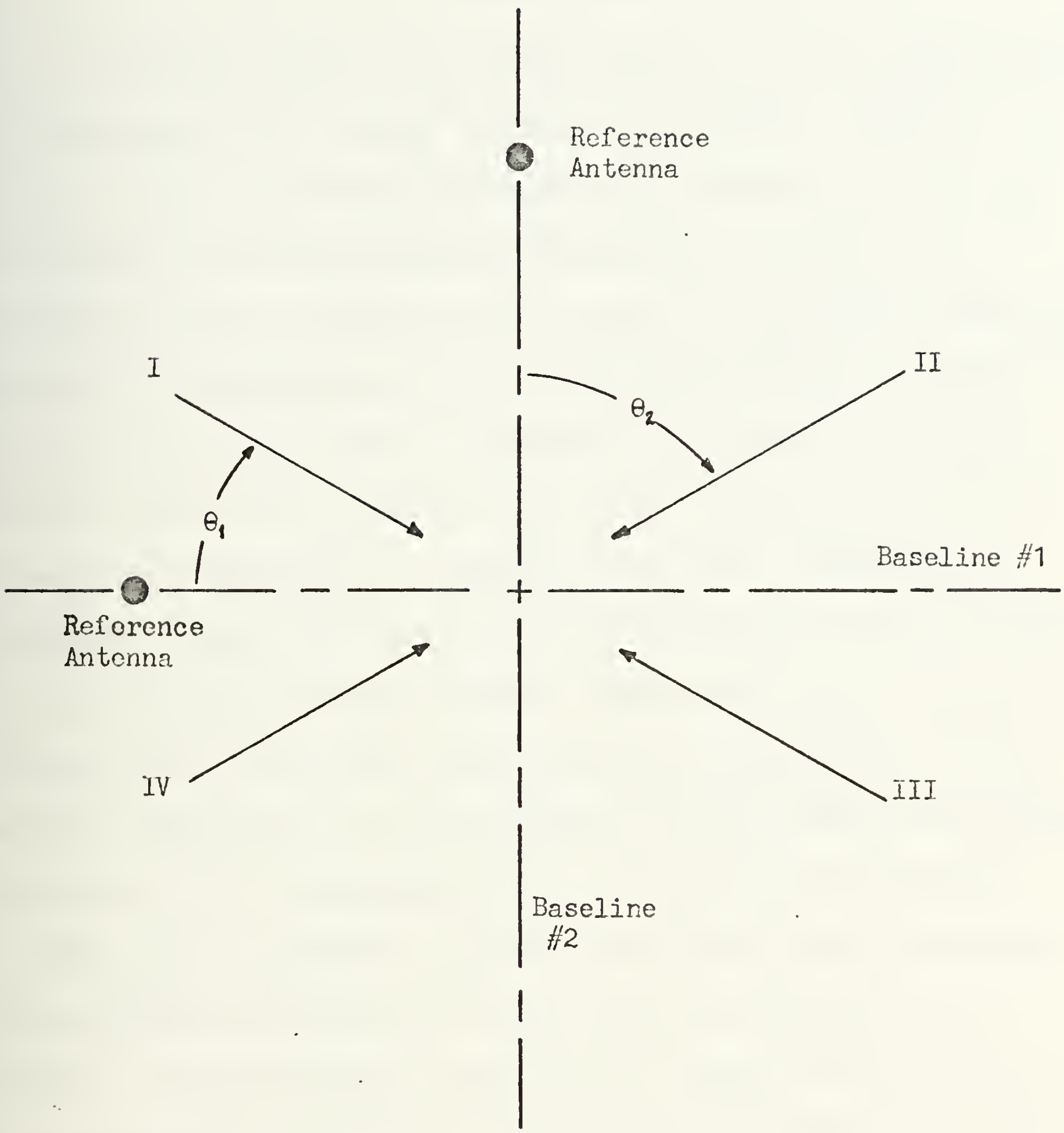


Figure 2-17, Two Array Interferometer

III. SYSTEM MODELING

In order to predict the effectiveness of the phase coded RDF technique in an actual signal environment, certain aspects of its performance should be evaluated. As in any RDF system, the most important parameters by which its usefulness can be judged are its angular resolution capability, and its performance in a noisy environment. It has already been shown that an idealized two array system is possible which can eliminate the gross bearing errors due to angular ambiguities. However, the basic performance parameters which will govern its precision are those inherent to the individual linear arrays. Specifically, the bearing accuracy and resolution capability of the system is fundamentally dependent on the precision of the signal processing technique used to determine time of arrival differences.

Therefore a computer simulation of the signal encoding and decoding portion of the RDF system was employed to evaluate its performance under noisy signal conditions, and to determine the effects of varying array lengths and signal parameters. The program was originally developed by R. C. Todaro in his Master's thesis [1] to test the uniqueness of the correlation outputs for varying degrees of correlation filter mismatch. The original signal model used in the simulation was an idalized rectangular RF pulse of fixed duration, and a noise-free environment was assumed. The output from this simulation consisted of printer plots of

autocorrelation and crosscorrelation outputs for various arrival angles, and for varying degrees of correlation filter mismatch.

The correlation results obtained from the original model indicated that no inherent ambiguities existed for arrival angles within any given quadrant of the array ground plane. In addition, the crosscorrelation plots obtained, using a 16-element array model, indicated that the angular resolution of the system was limited only by the precision of the matched filter delay units. That is, if delay lines could be physically realized so that their time delays corresponded to each incremental degree of arrival angle, then the resolution of the system would be $\pm 0.5°$.

A. MODEL DESCRIPTION

For the purpose of this paper, the program was modified to produce continuous fine-grain autocorrelation and cross-correlation output plots from a CALCOMP plotter for varying noise levels, array lengths, and input signal parameters. The system model is identical to the graphical illustration used in the previous chapter. A single linear array was used, with the azimuthal angle of arrival measured from the baseline of the array. The signal model was modified as shown in Figure 3-1, to investigate the effects of input signal-to-noise ratio on the correlation outputs. A constant noise level, corresponding to the average (or RMS) level in an actual signal environment, was chosen.

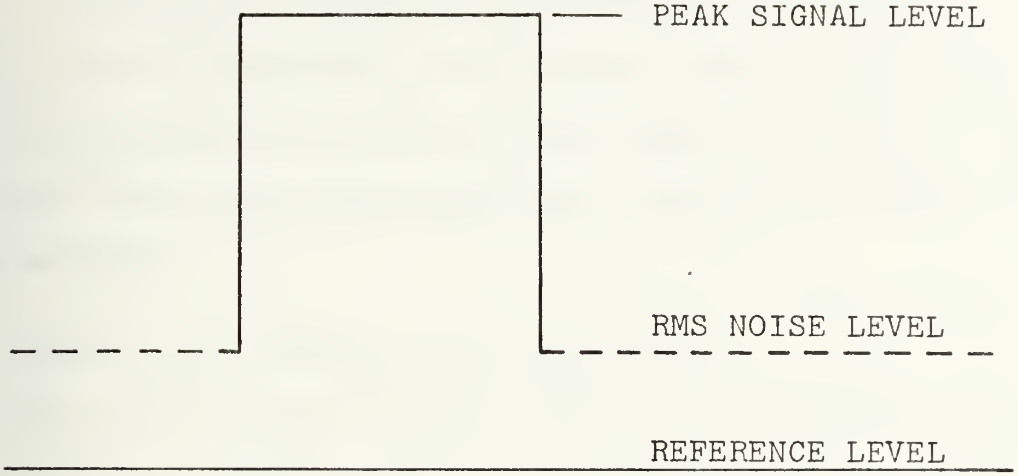


Figure 3-1. Idealized Signal Model

The signal-to-noise ratio was defined as the ratio of the peak signal voltage to the RMS noise voltage.

The model parameters that are variable, either by choice of specific program data cards, or by internal adjustments to program instructions, are as follows:

1. Array length (2,4,8,10,16, and 18 may be selected by data cards, larger arrays require re-dimensioning of program storage).
2. Arrival angle (selected by data cards)
3. Input signal-to-noise ratio (selected by two program instructions).
4. Choice of autocorrelation or crosscorrelation outputs (selected by data cards).
5. Complementary sequences (any valid complementary code pair may be loaded via data cards, the filter codes for autocorrelation or crosscorrelation processes are automatically derived by the program).

6. Filter/Signal matching coefficient, D (the ratio of individual correlation filter delay to maximum propagation delay Δt, selected by data card).
7. Input pulse width (selected by program instruction).

A detailed discussion of the program operation, format of input data, and an instruction listing are included in the appendix.

B. MODEL OUTPUT RESULTS AND CONCLUSIONS

The output plots obtained from this simulation are divided into three specific groups in order to show the effects of individual parameter changes on the overall operation of the model. These groups are outlined below, and are referenced to specific figure numbers, so that the related discussion and conclusions may be presented together at the end of this section. All pertinent parameters are listed in the heading of each of the plots.

1. Group 1, Figures 3-2 through 3-41

These plots demonstrate the effect of varying the input signal-to-noise ratio on autocorrelation and cross-correlation outputs. SNR takes on values of infinity (no noise), 10:1, 5:1, and 2:1.

Fixed parameters: Arrival Angle = 60.0°

Array Length = 16

Pulse Duration = 2.0 times maximum inter-antenna propagation time.

For each combination of SNR and fixed parameters the matching coefficient (D) takes on values of 0.900, 0.950, 1.000, 1.050 and 1.100.

2. <u>Group 2, Figures 3-42 through 3-56</u>

These plots demonstrate the effect of varying array length on autocorrelation and crosscorrelation outputs. Array lengths of 10, 8, and 4 elements are utilized.

Fixed parameters: Arrival angle = 60.0°

Pulse Duration = 2.0 times maximum inter-antenna propagation time.

For each combination of array length and fixed parameters, the matching coefficient takes on values of 0.900, 0.950, 1.00, 1.050 and 1.100.

3. <u>Group 3, Figures 3-57 through 3-80</u>

Effect of varying pulse width/antenna separation ratio on autocorrelation and crosscorrelation outputs. (Pulse durations of 20 and 200 times the inter-antenna propagation time are utilized).

Fixed parameters: Arrival Angle = 60.0°

SNR = Infinite (no noise)

4. <u>Discussion and Conclusions</u>

The plots comprising Group 1 are autocorrelation and crosscorrelation outputs for four decreasing values of signal-to-noise ratio. The first series, shown in Figures 3-2 through 3-11, is for an ideal noise-free signal environment. The progression of the matching coefficient from 0.900 to 1.100 in increments of 0.05 shows the expected result for the perfectly matched case (D = 1.000). The autocorrelation output is an exact replica of the input

pulse with a gain factor of 32, or twice the array length. The crosscorrelation output is everywhere zero for the matched case.

As the signal-to-noise ratio begins to take on finite values, however, the crosscorrelation outputs no longer exhibit a zero value for the matched case. It is also evident that the autocorrelation outputs exhibit finite signal-to-noise ratios in the presence of input noise. A comparison of the matched autocorrelation outputs for input signal-to-noise ratios of 10:1, 5:1, and 2:1 shows that the output signal-to-noise ratios are numerically equal to the input signal-to-noise ratios.

The most important conclusion to be reached from this first group of plots, is that the crosscorrelation outputs still possess unique characteristics that distinguish them from the outputs of the mismatched filters, even in the presence of input noise. These characteristics are minimum amplitude and unipolar waveforms. The peak amplitudes of the crosscorrelation output waveforms show marked reductions as the perfectly matched case is reached. This characteristic is evident in the plots in Group 2 for array lengths of 10, 8, and 4 as well. In general, for an input signal-to-noise ratio of 5:1, all crosscorrelation outputs show an approximate 300% reduction in peak output as the filter characteristic progresses from a mismatched condition to a perfectly matched condition (regardless of array length).

The second identifying feature of the crosscorrelation outputs is their polarity. For the mismatched case the outputs exhibit large amplitude noise peaks of both positive and negative polarity. When the filters provide exact matching, however, the output is of a single polarity. For the 16-element array utilized in Group 1 these unipolar outputs are on the positive side. An interested quality of these unipolar outputs was noticed during the plotting of the various outputs for this paper. That is, the positive or negative polarity of these perfect match crosscorrelation outputs can be reversed by merely reversing the order of the complementary sequences used to phase code the input signals. While no attempt is made here to justify this result, it is mentioned because it might be of value in the design of recognizers for scanning the various correlation filter outputs.

It should also be noted that the matched crosscorrelation outputs do not possess unipolar qualities for all array lengths. For instance, the matched crosscorrelation output (with input noise) for an array length of 10, as shown in Figure 3-44, is bipolar. It was assumed, therefore, that this characteristic is dependent upon the chosen code length. Not enough data is available here to attempt any further explanation of this characteristic.

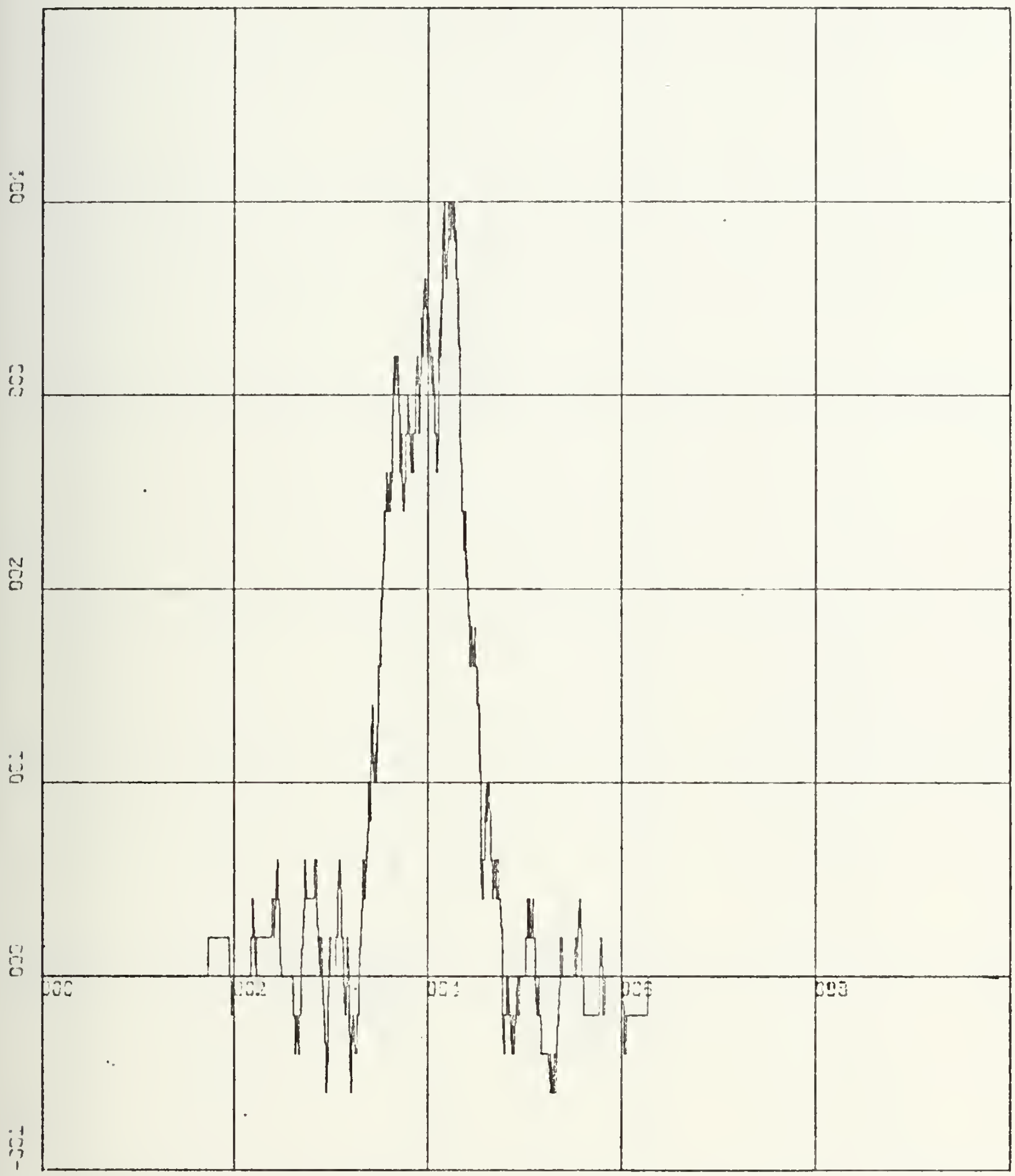
The plots comprising Group 2 are autocorrelation and crosscorrelation outputs for array lengths of 10, 8, and 4 elements. These plots were obtained in order to determine

the effect of array length on the resolution capability of the system. These plots, and those included in Group 1, can be used to derive an approximate maximum value of the resolution. That is, for matching coefficients of 0.950 and 1.050, which correspond to filters matched to arrival angles 1.7° above and below the true arrival angle, the correlation outputs are indicative of a mismatched condition. Therefore, the resolution of the system is at most 1.7°. Although finer resolutions are probable, and more precise limits could be assigned to this parameter by comparison of additional output plots, it is important to note that the resolution is not degraded by the shorter array lengths. The basic conclusion reached from these results was that a system utilizing short (2 or 4 element) arrays is technically feasible. It would seem that the only advantage to be derived from using longer array lengths is the increased signal processing gain in the autocorrelation mode.

The plots comprising Group 3 are autocorrelation and crosscorrelation outputs for pulse widths normalized to 20 and 200 times the maximum inter-antenna propagation time. This propagation time is equivalent to the time difference between antenna responses for a wavefront incident at an arrival angle of zero degrees. It is obvious from these plots that the resolution of the system is degraded for signals of increased duration. The crosscorrelation outputs which were previously characteristic of the perfectly matched filter, now result for a range of filter mismatch.

For 20X pulse durations this output occurs over a range of filter mismatches from D = 1.000 (60° match) up to D = 1.050, which corresponds to an arrival angle of 58.3°. For the 200X pulse this range of ambiguous outputs runs from 60°, for a normally perfect match, up to a matching coefficient corresponding to 25.8°. This last series of output plots indicates that the proposed system would not be effective against signals with long duration modulation events.

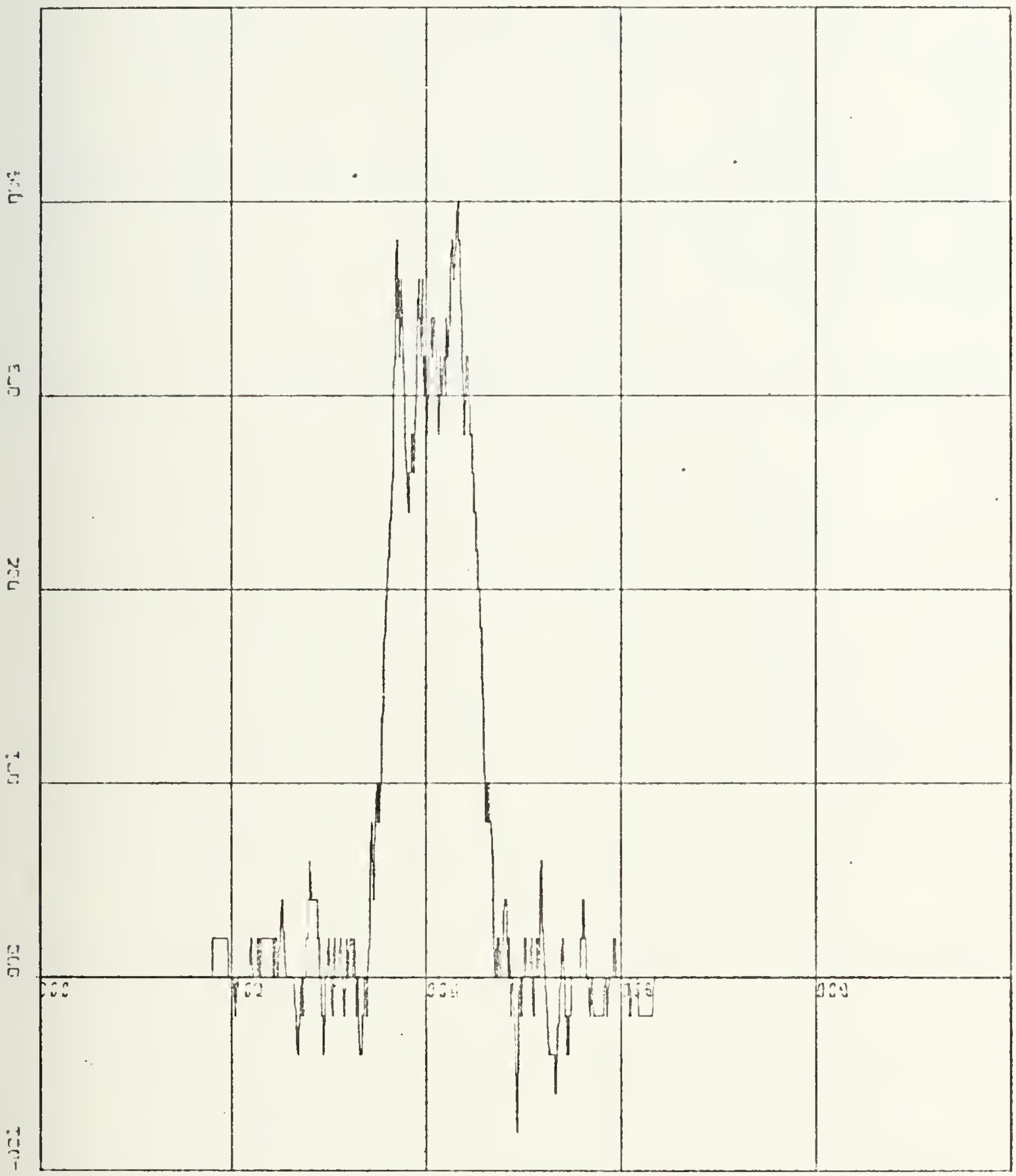
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-2

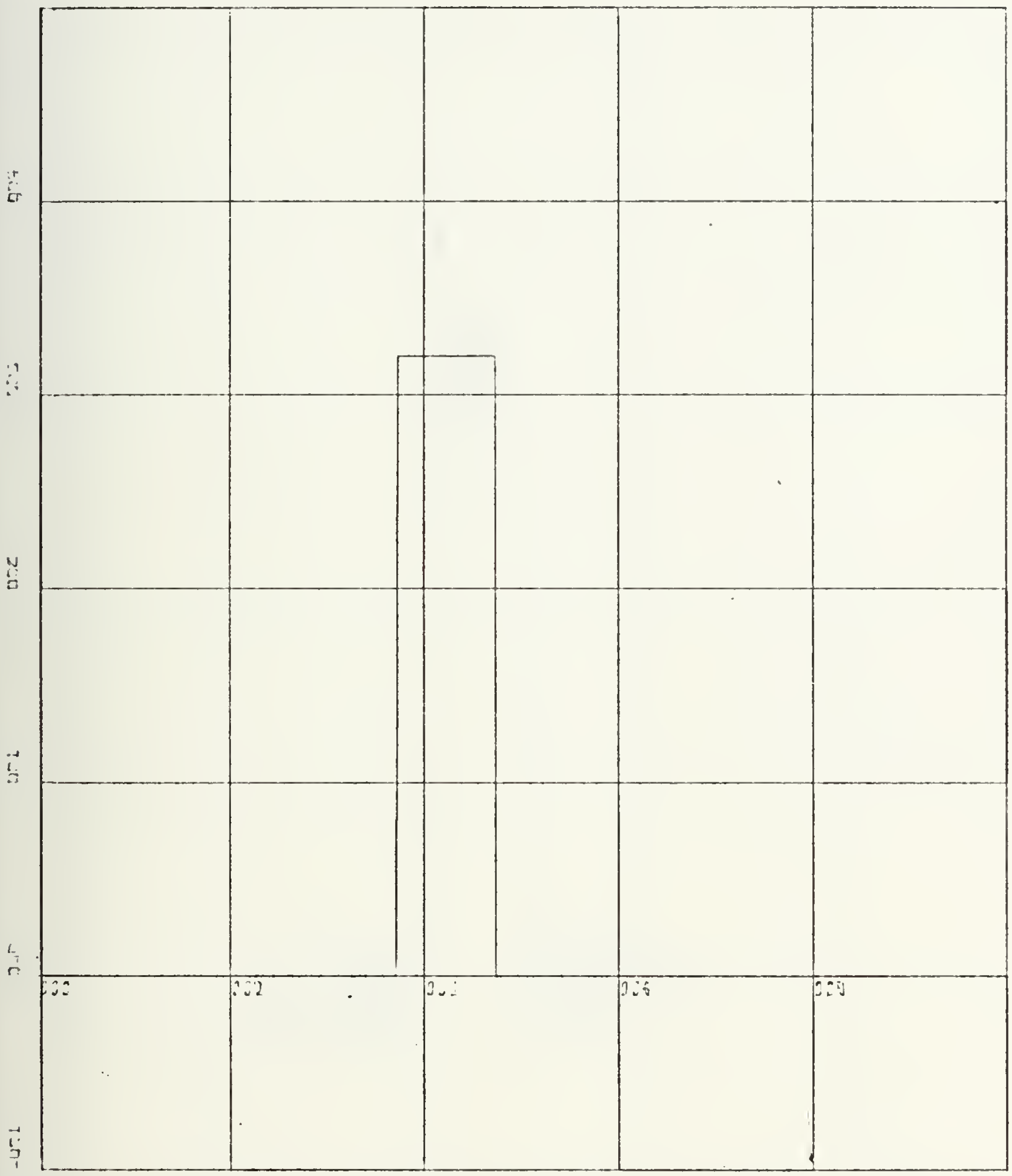
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-3

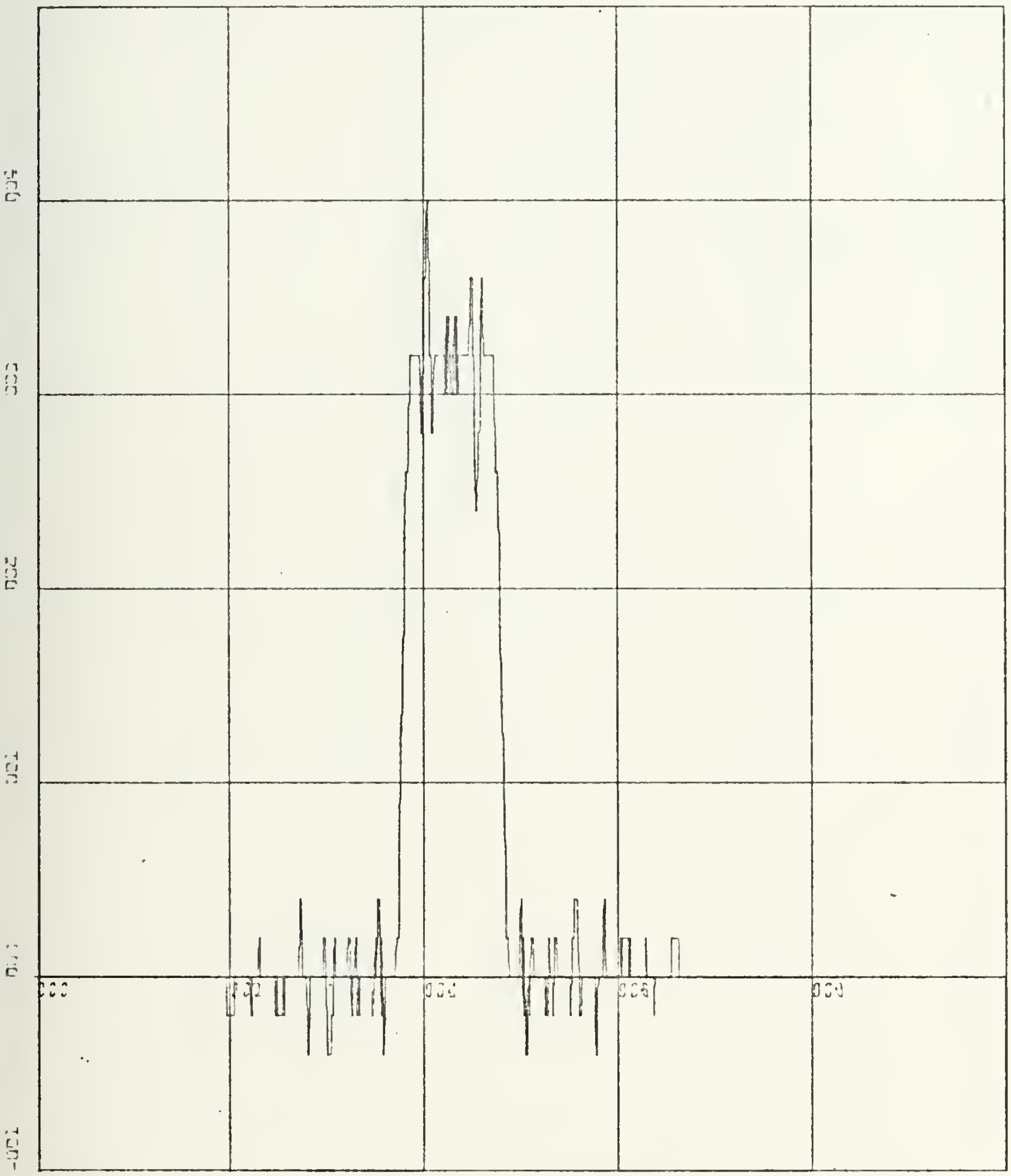
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-4

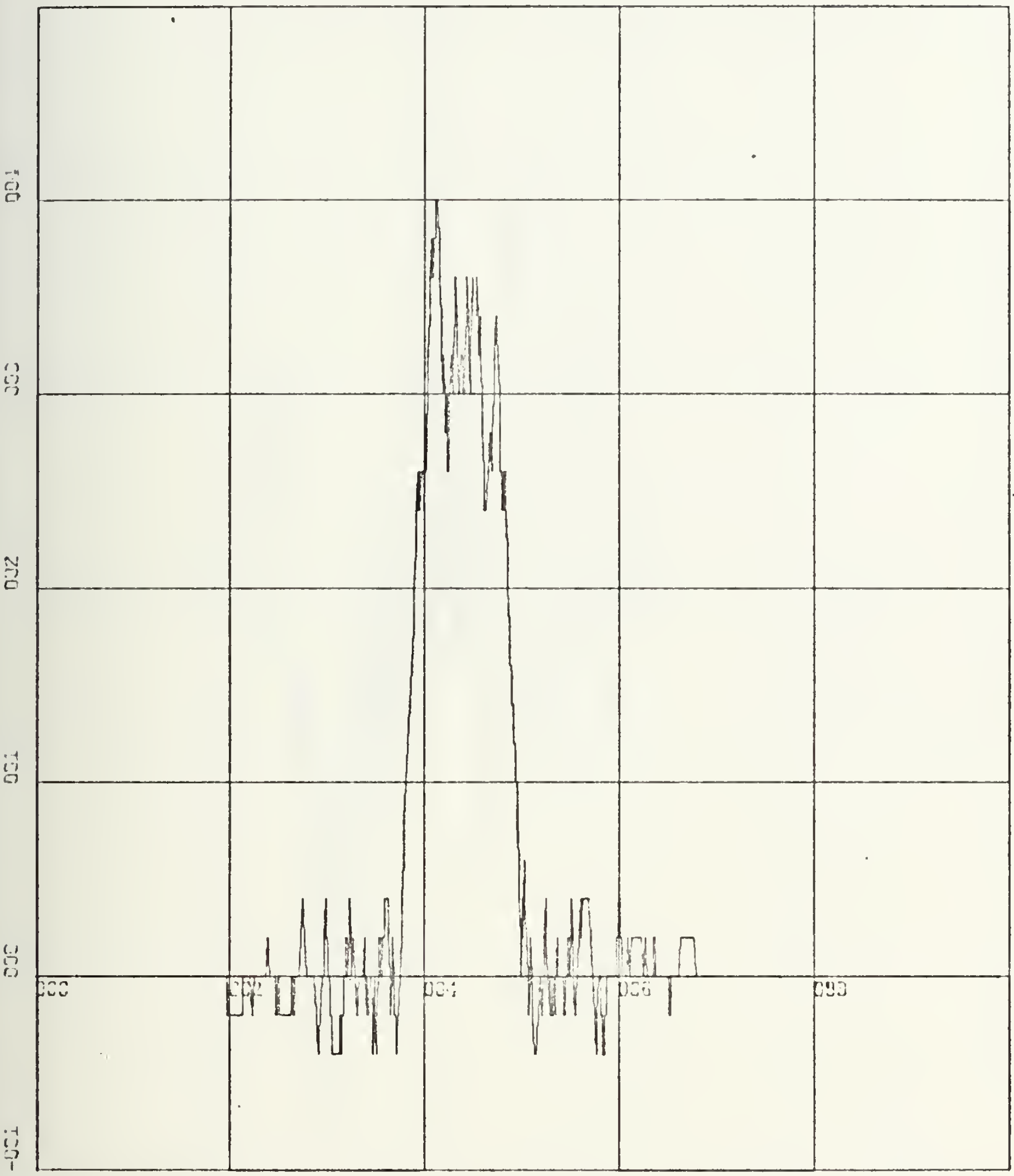
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-5

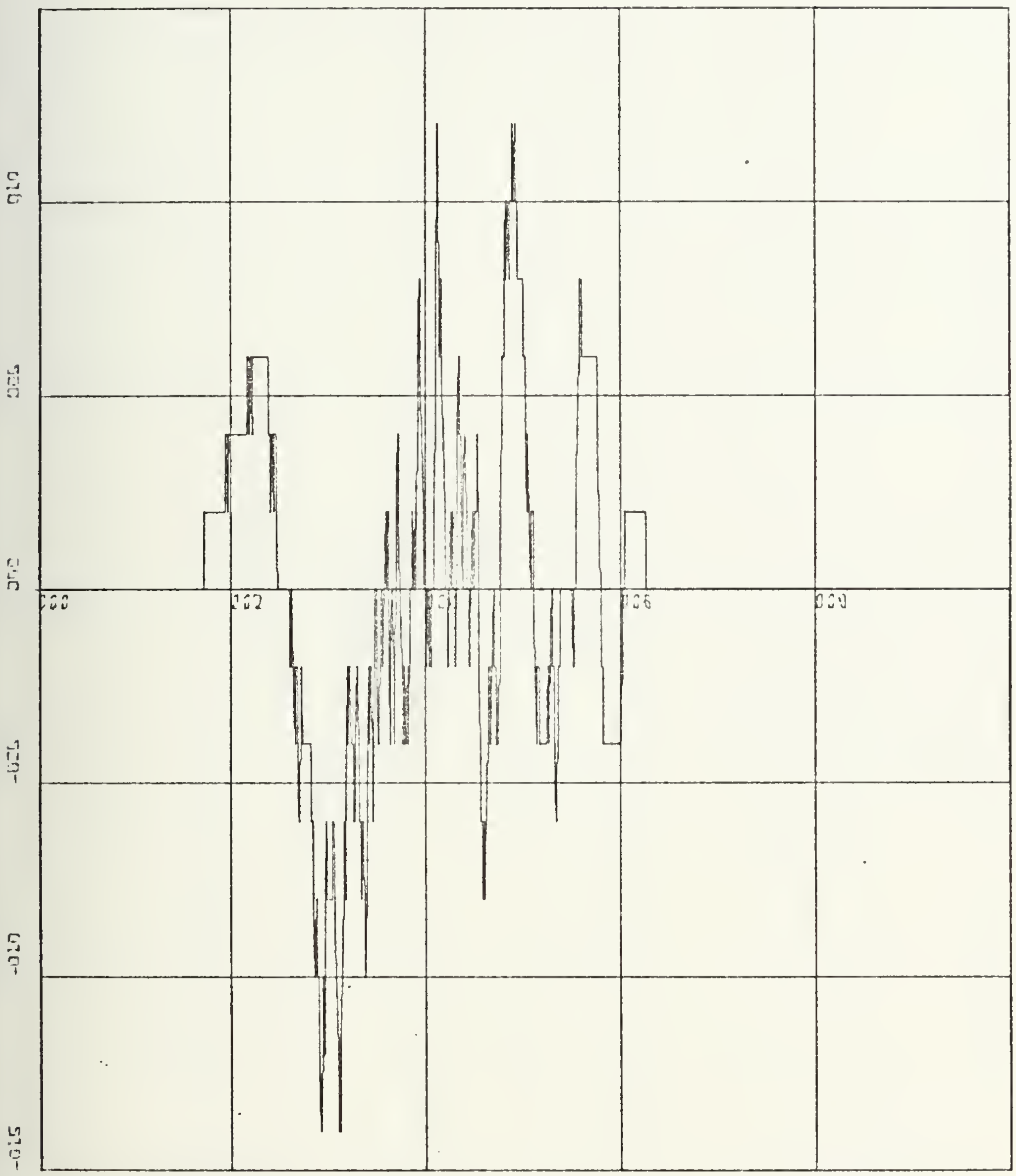
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-6

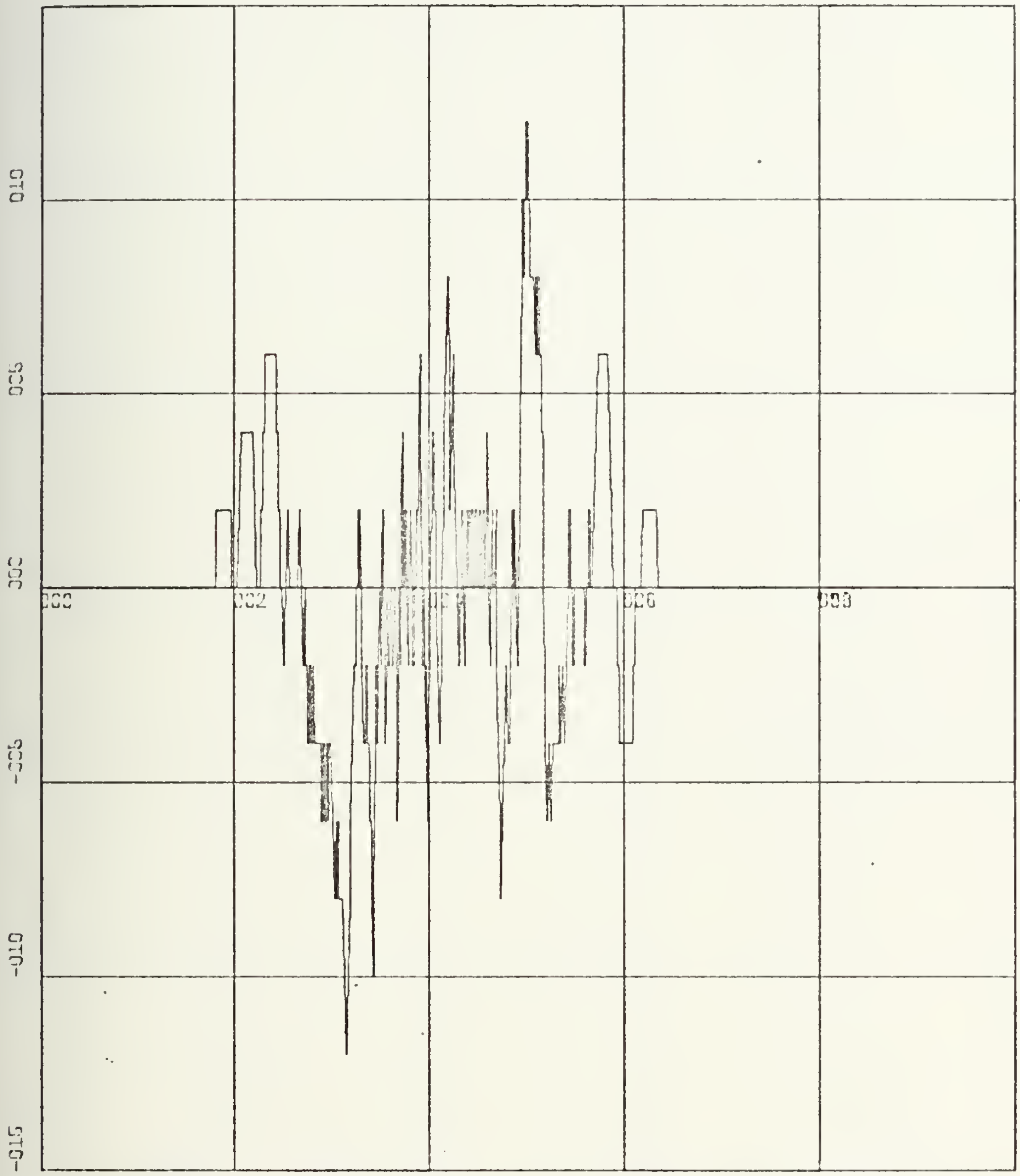
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-7

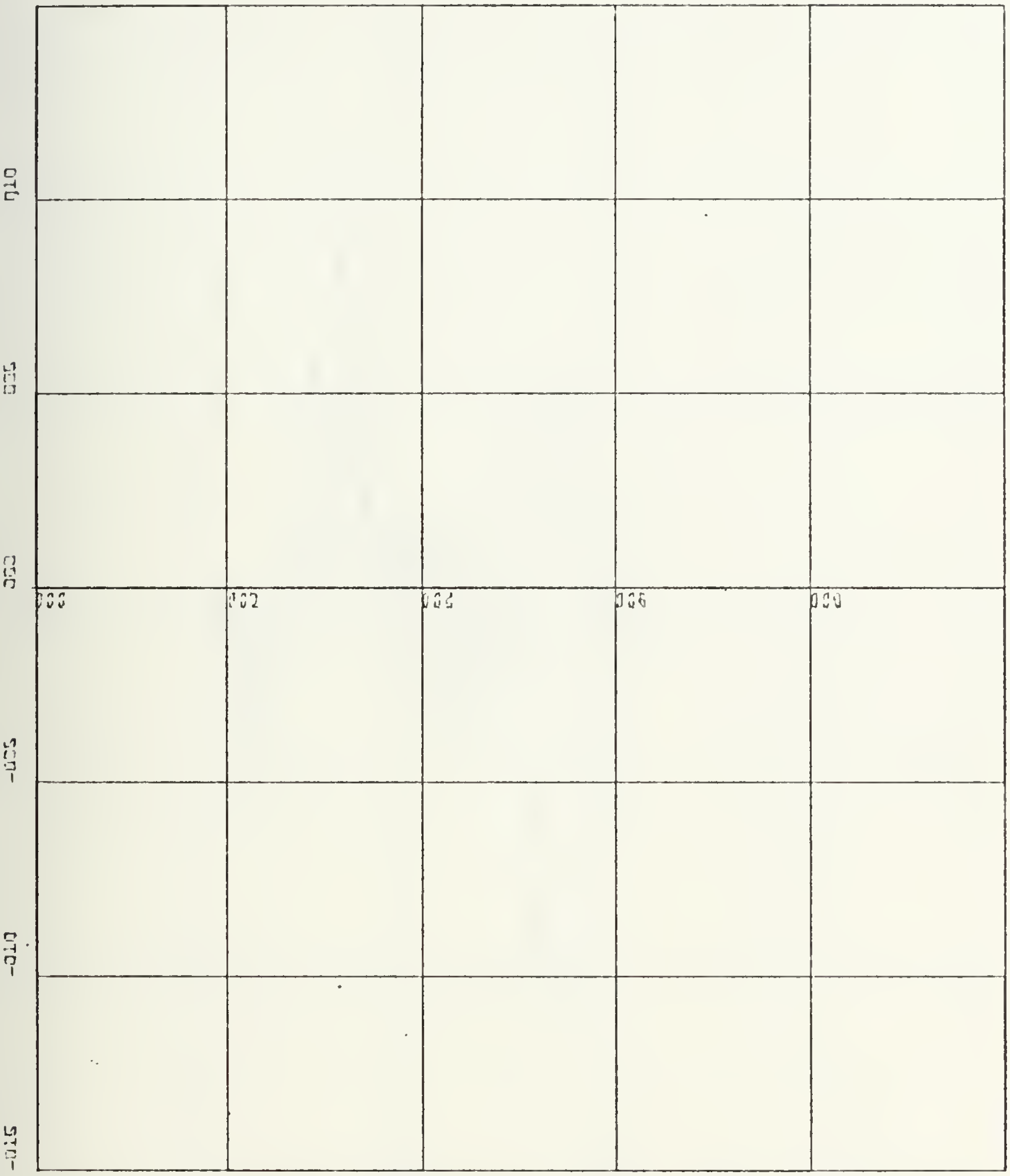
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°) ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-8

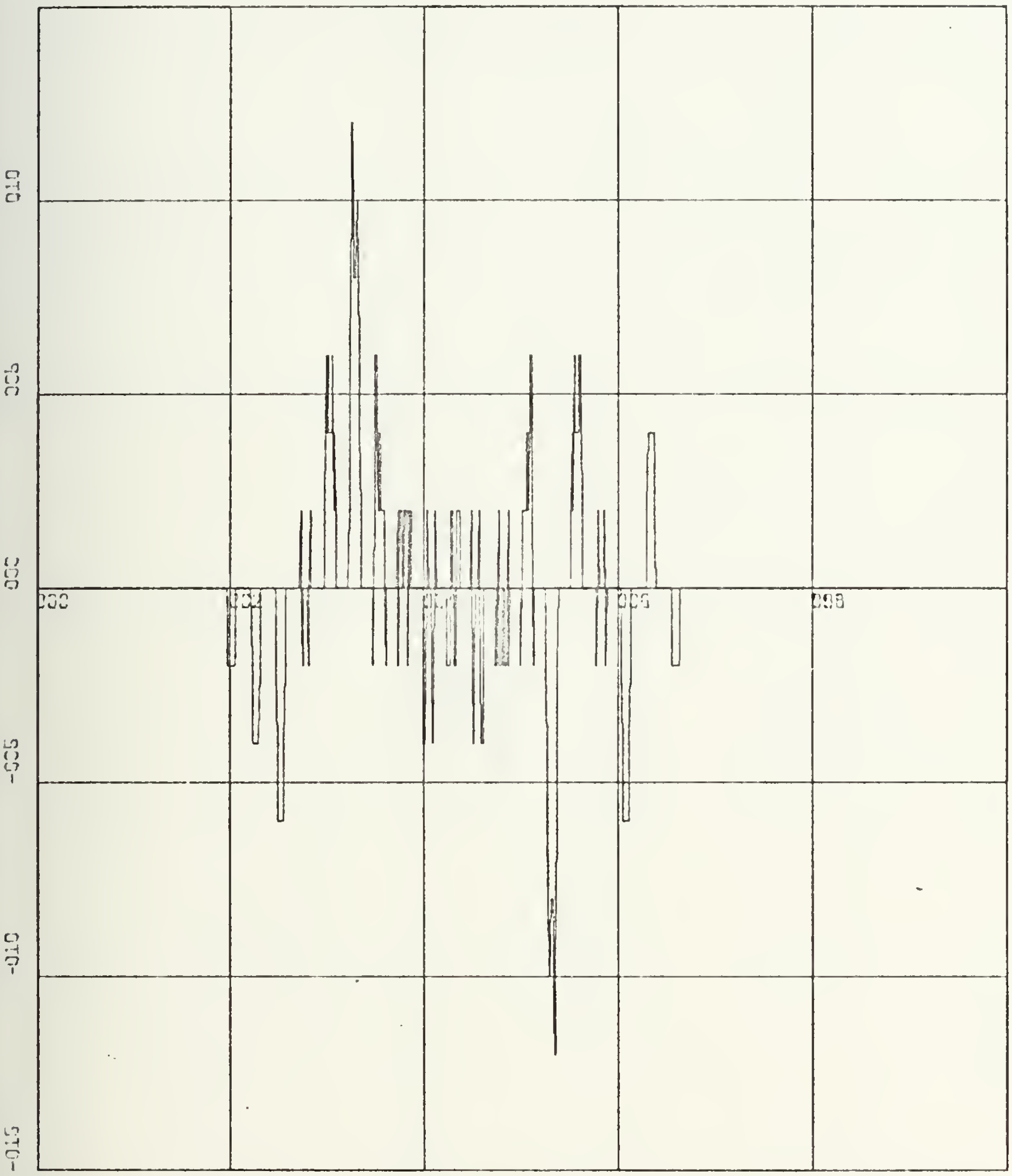
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-9

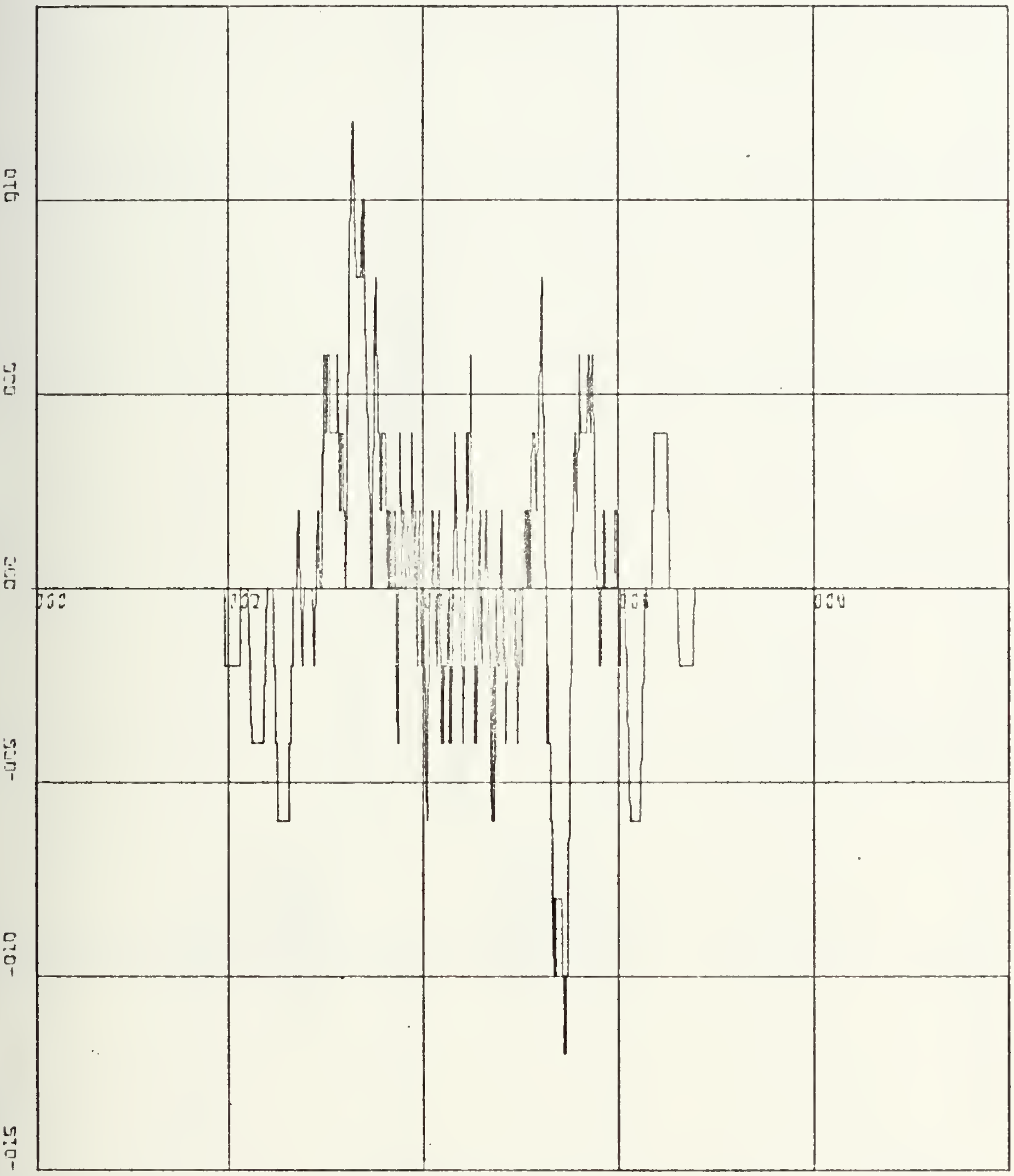
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-10

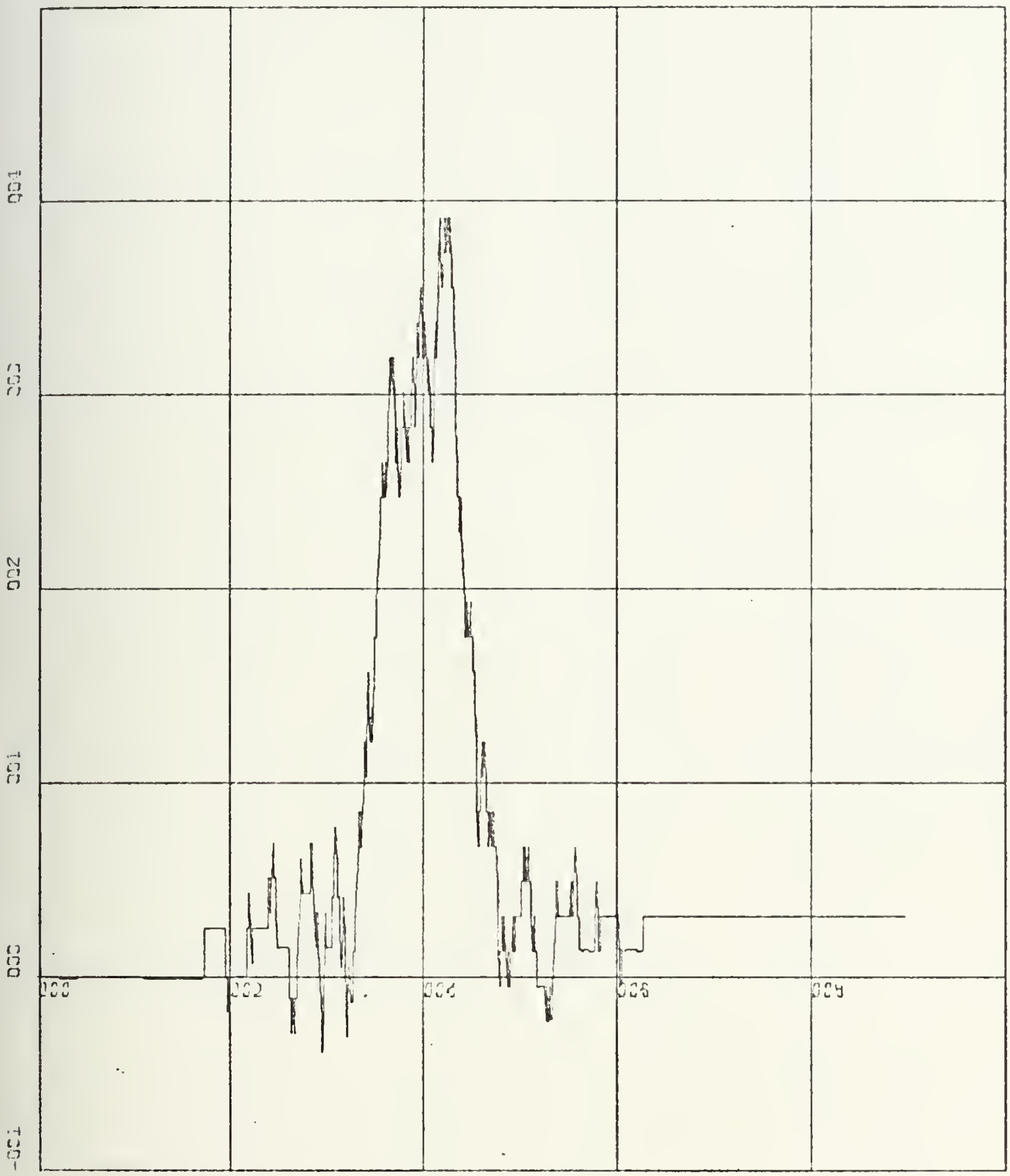
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-11

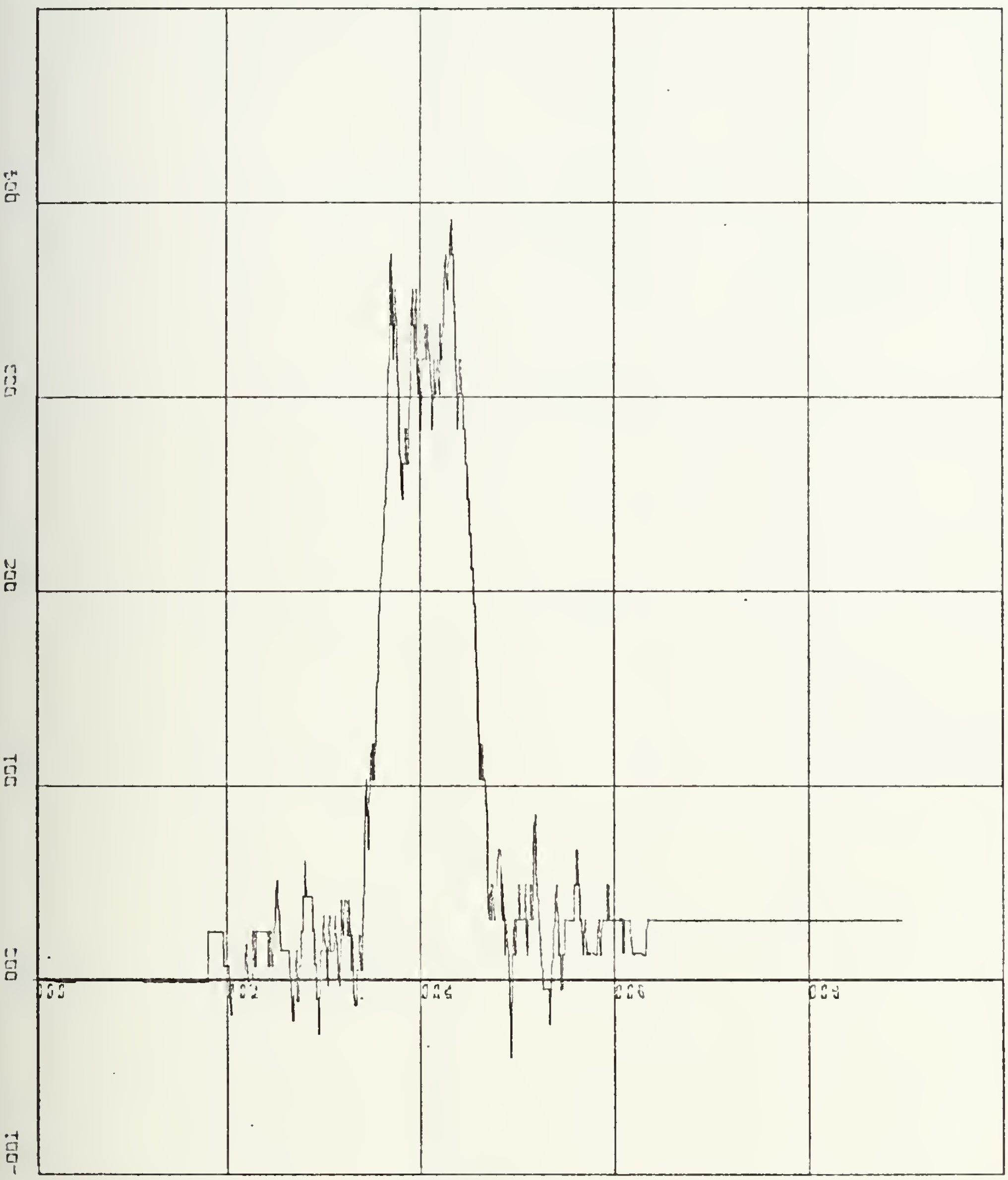
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-12

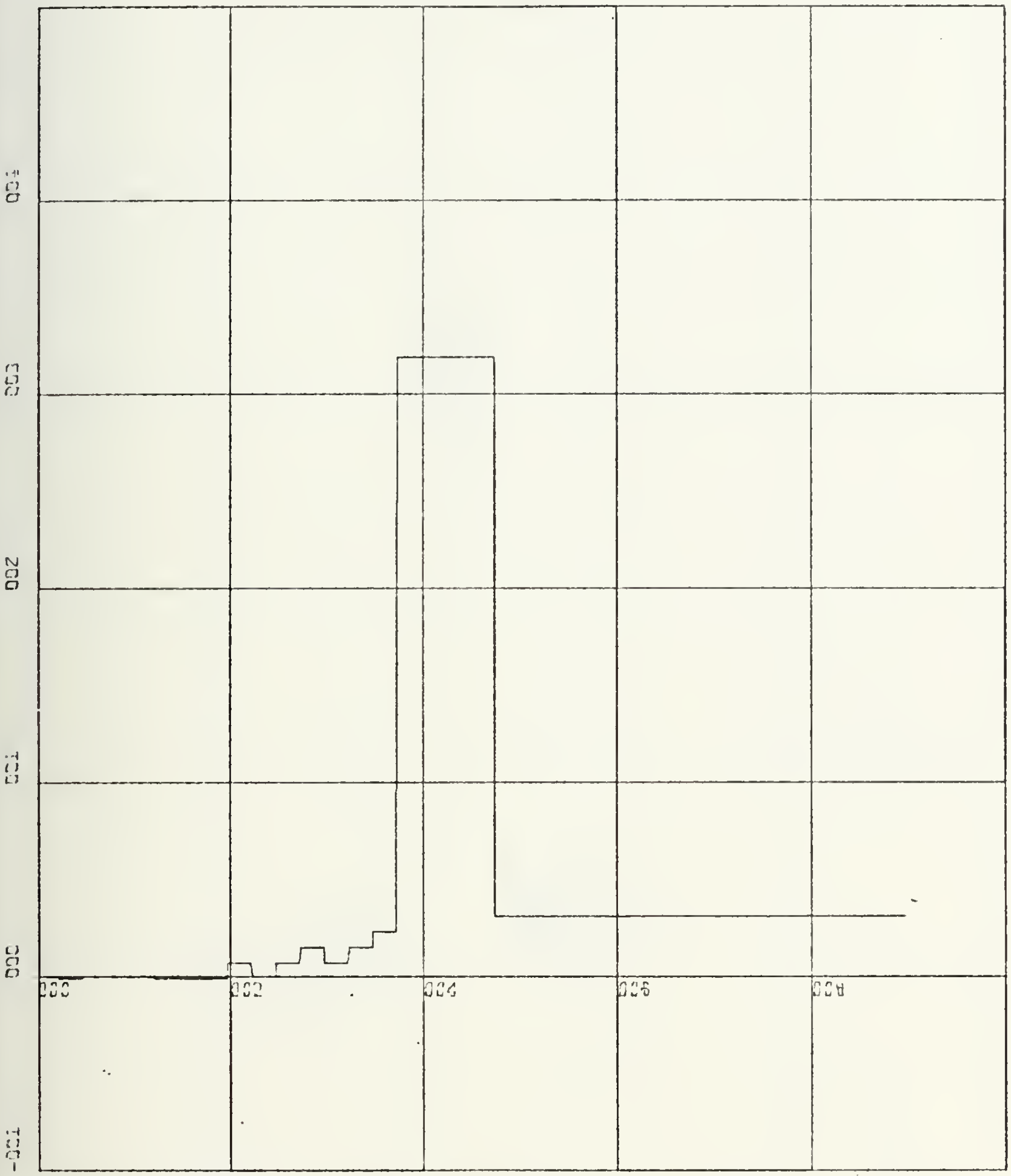
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-13

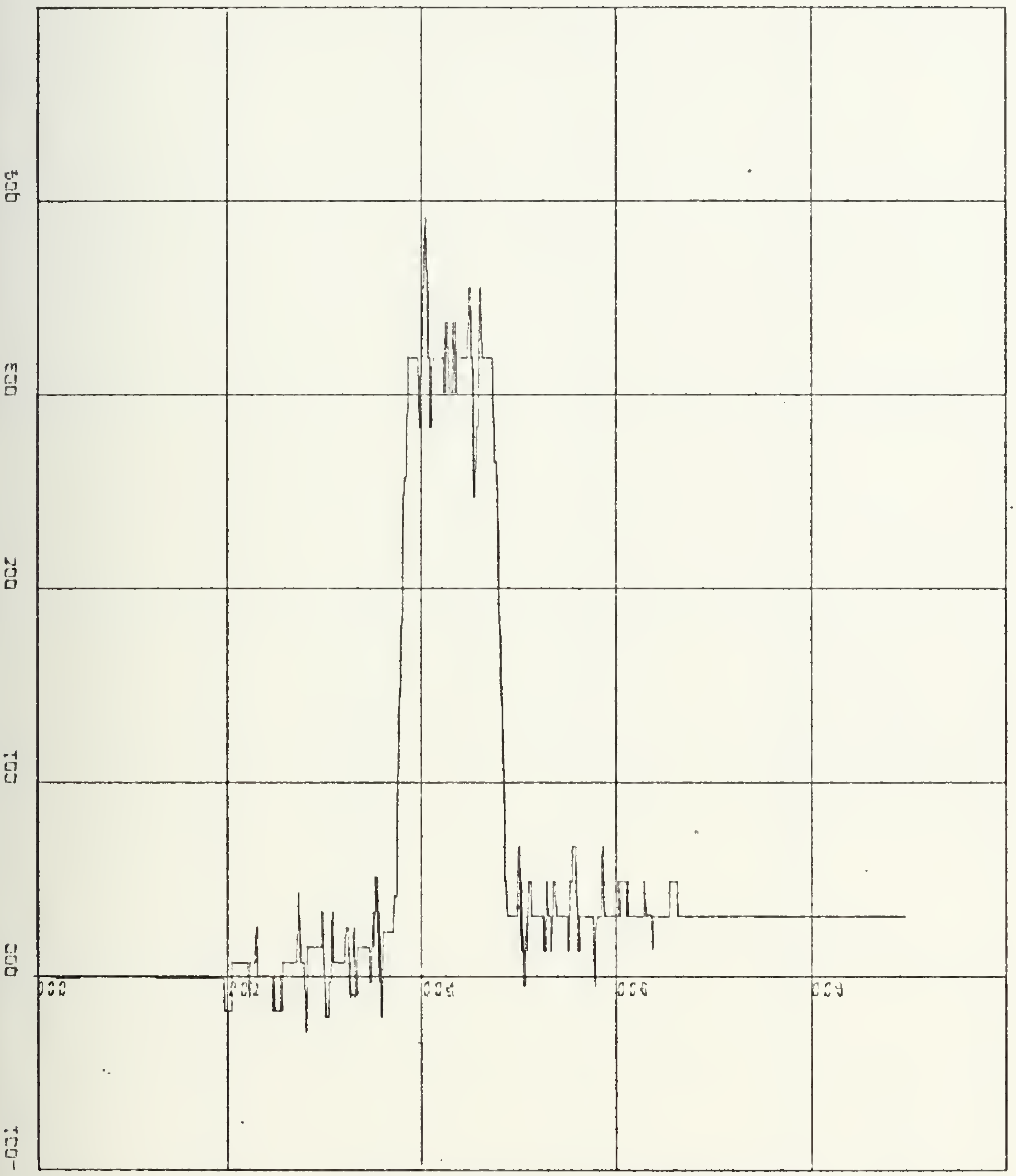
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-14

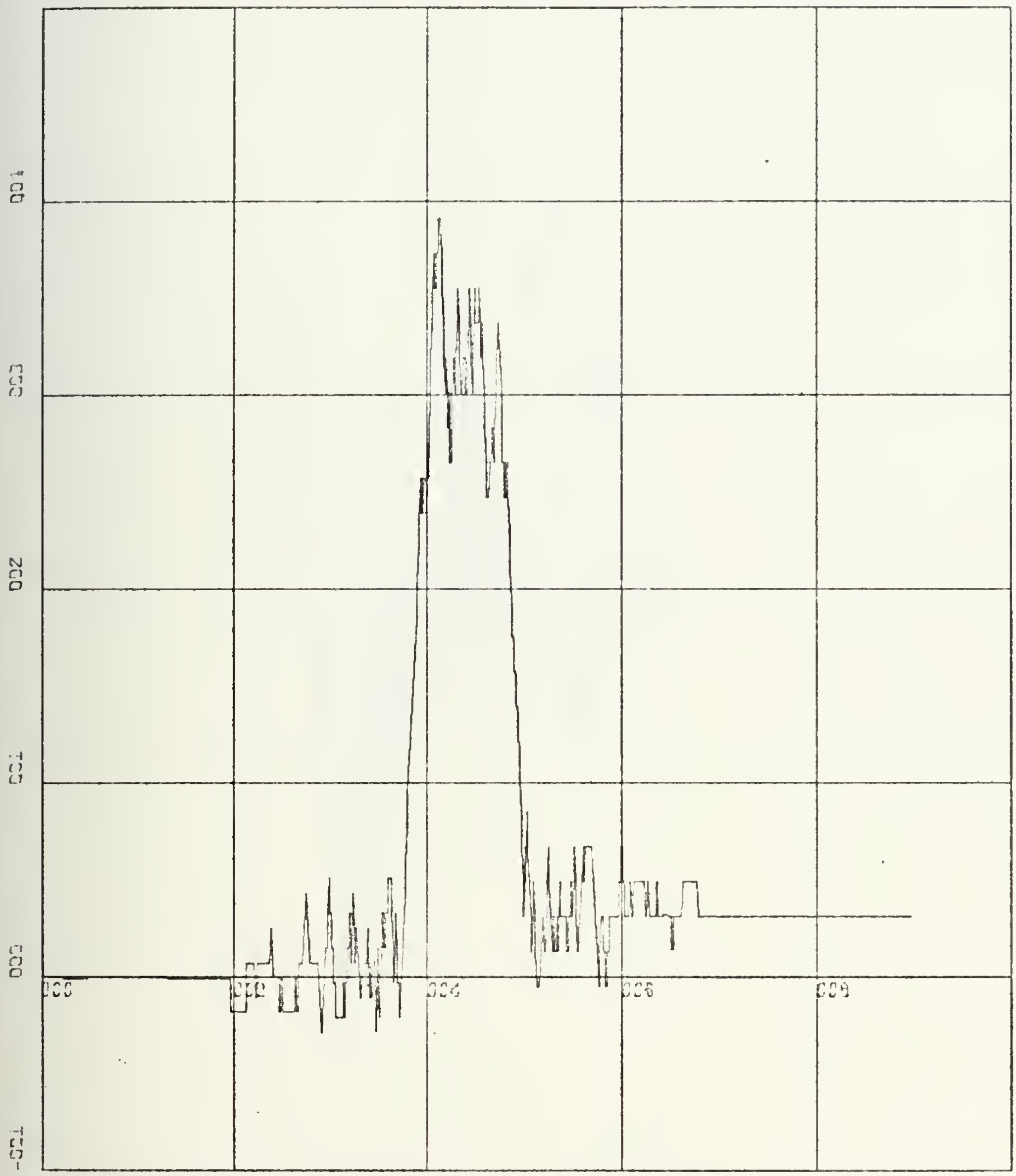
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-15

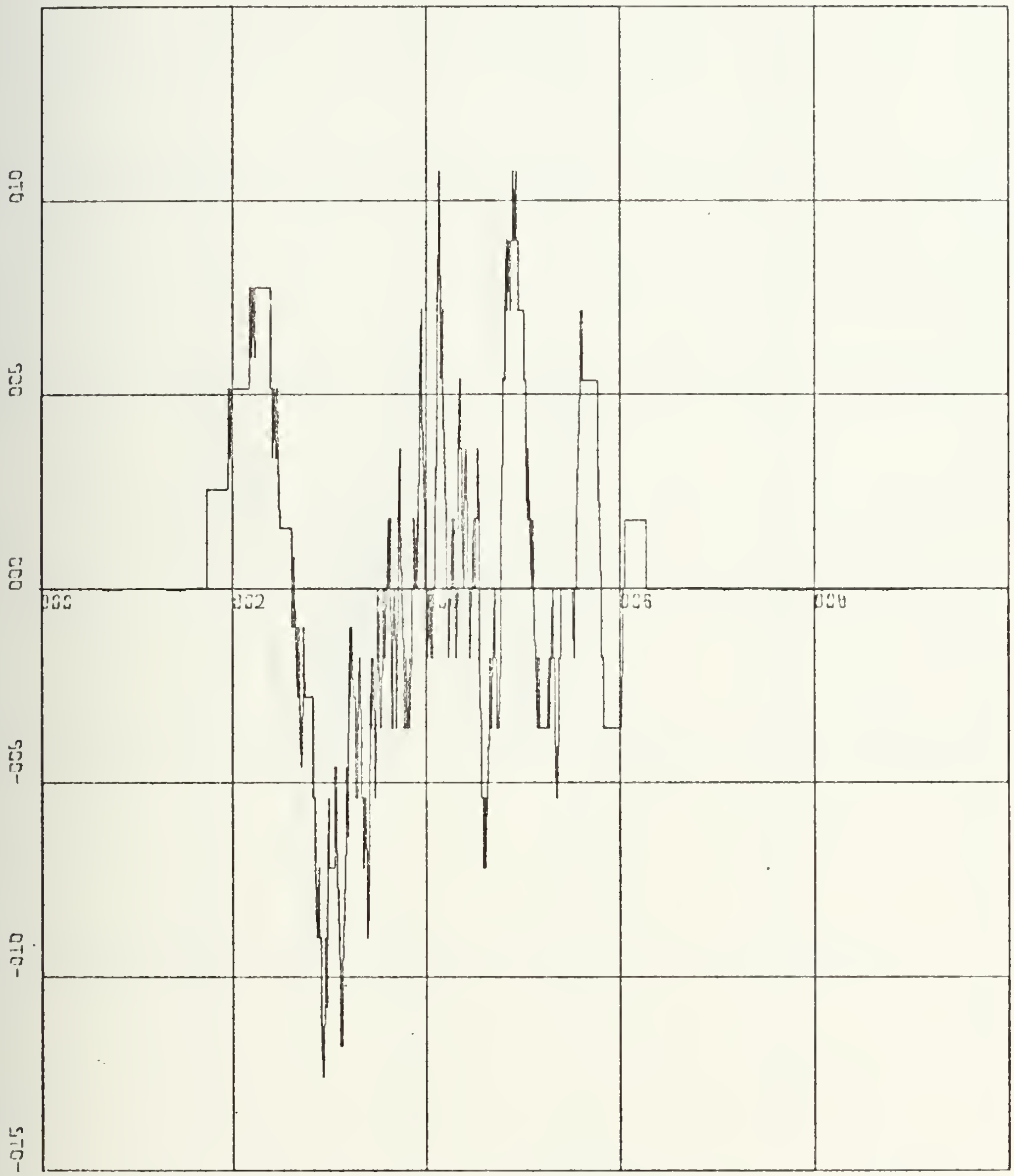
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-16

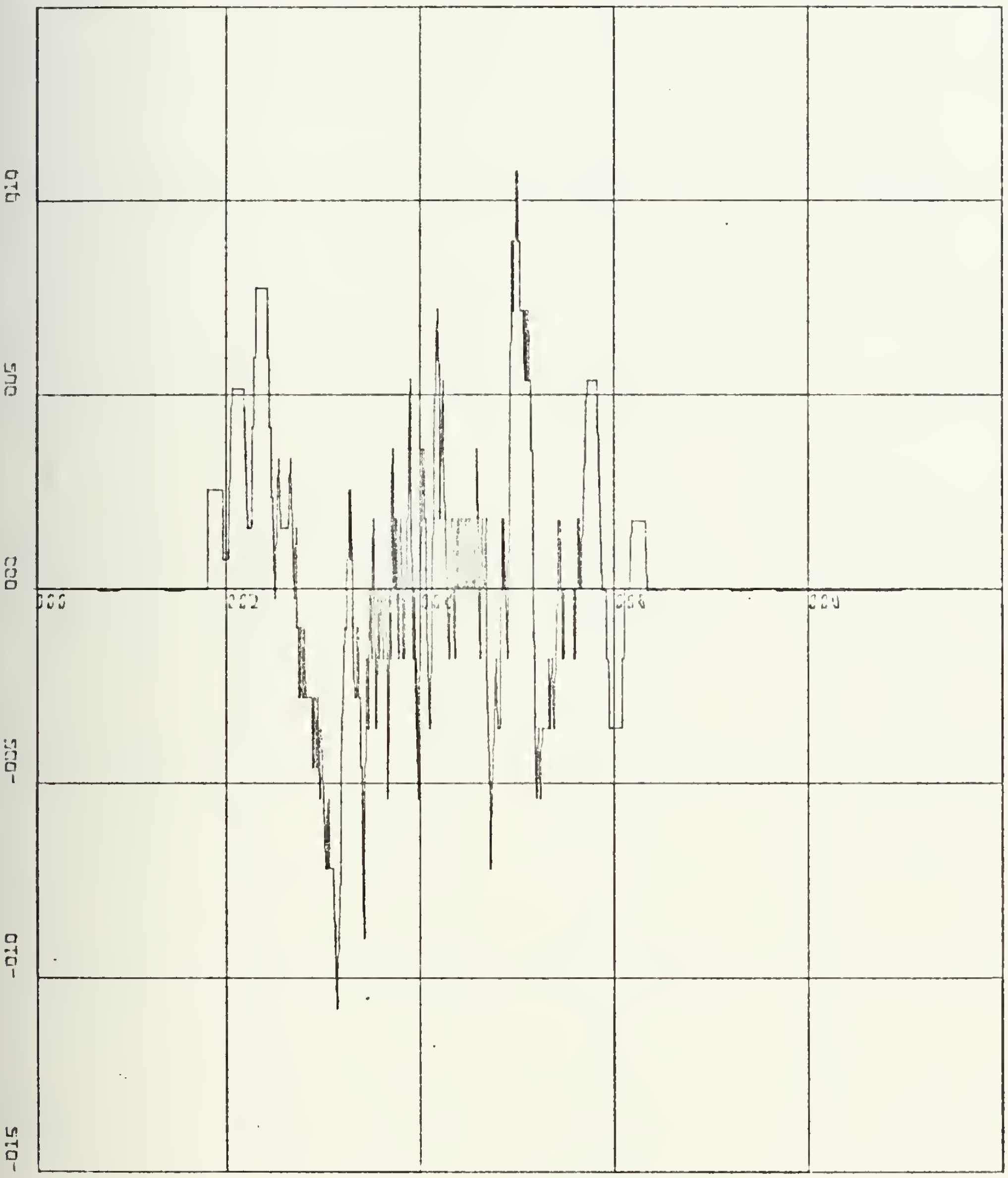
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY,



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-17

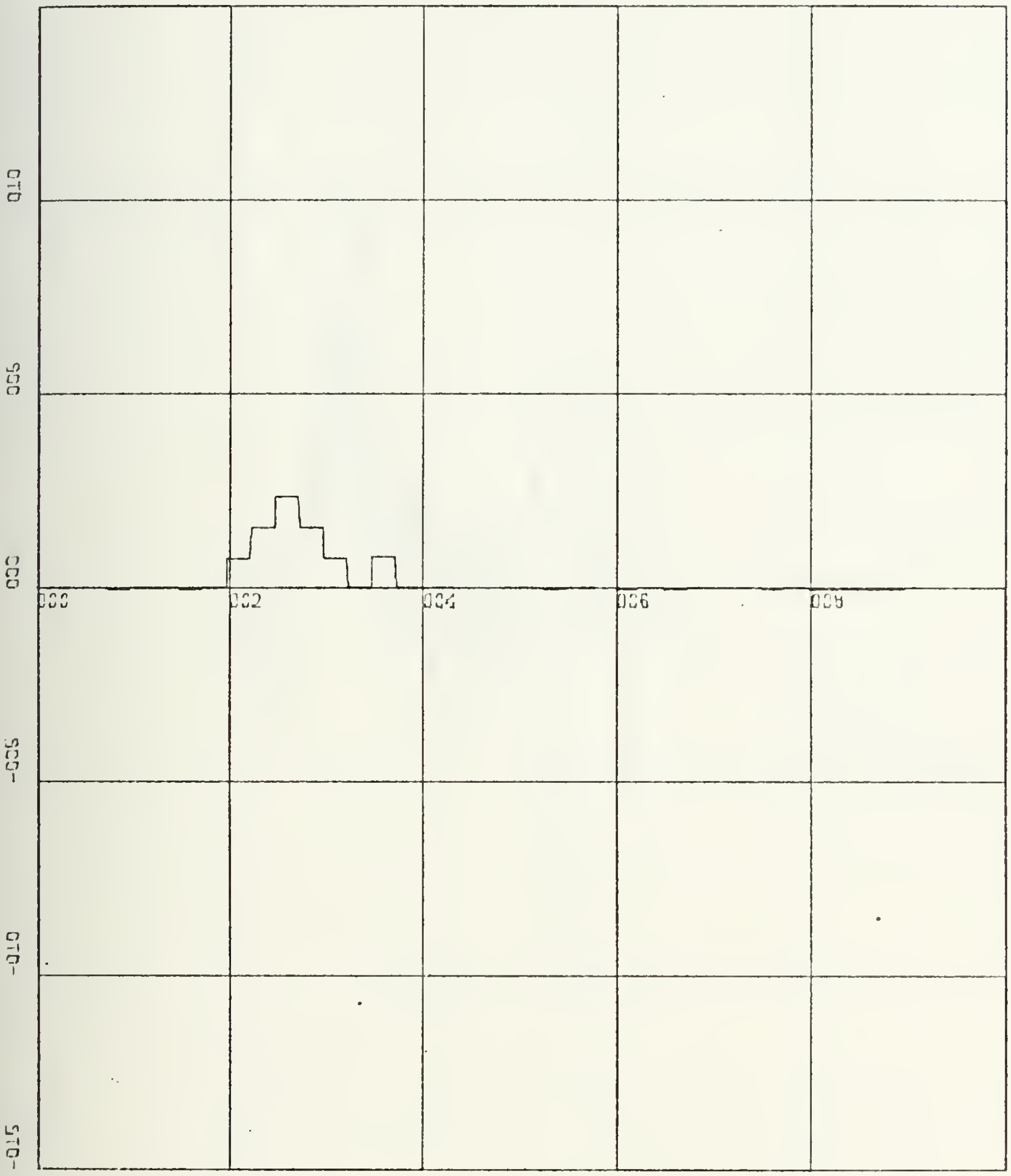
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-18

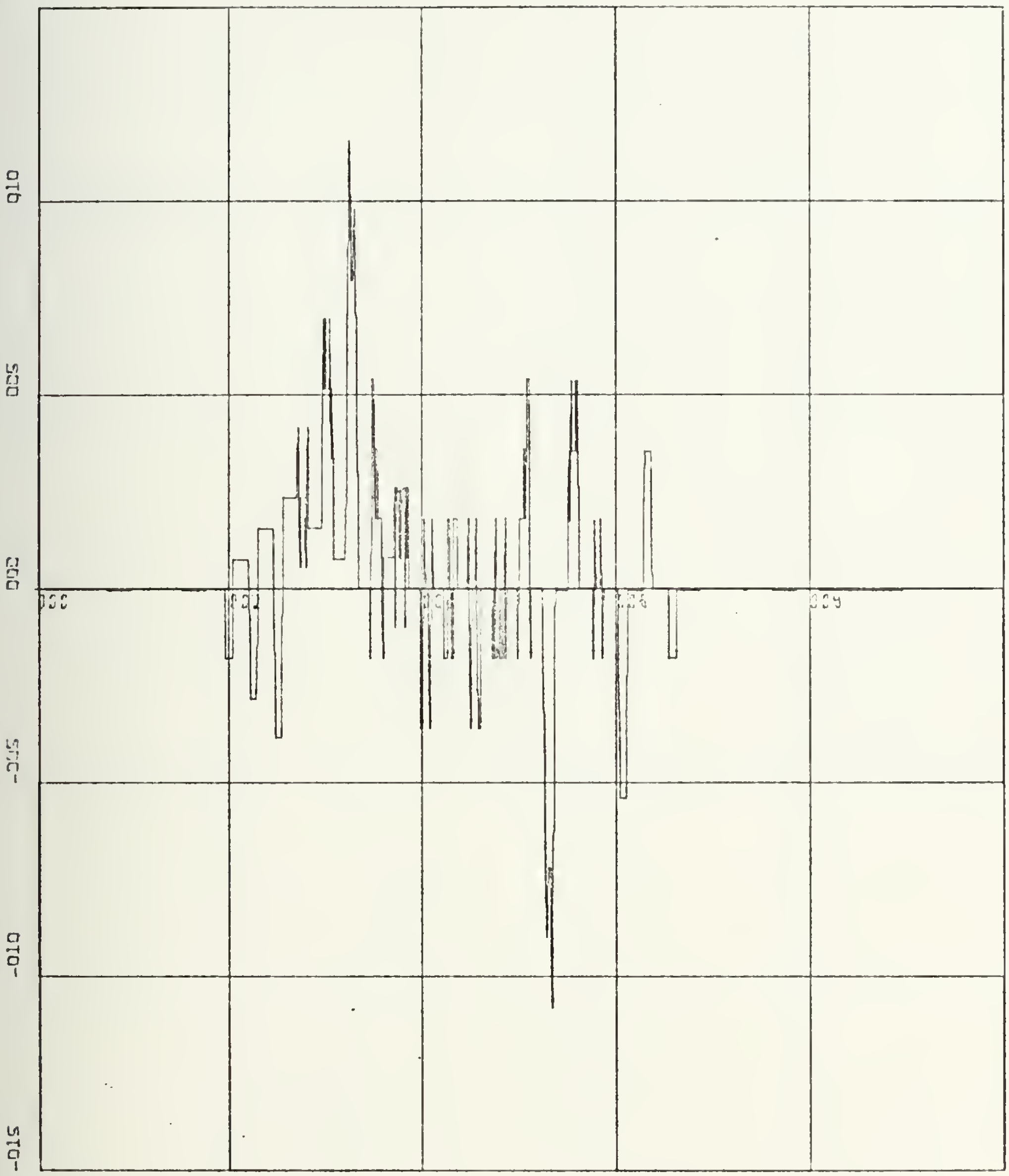
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-19

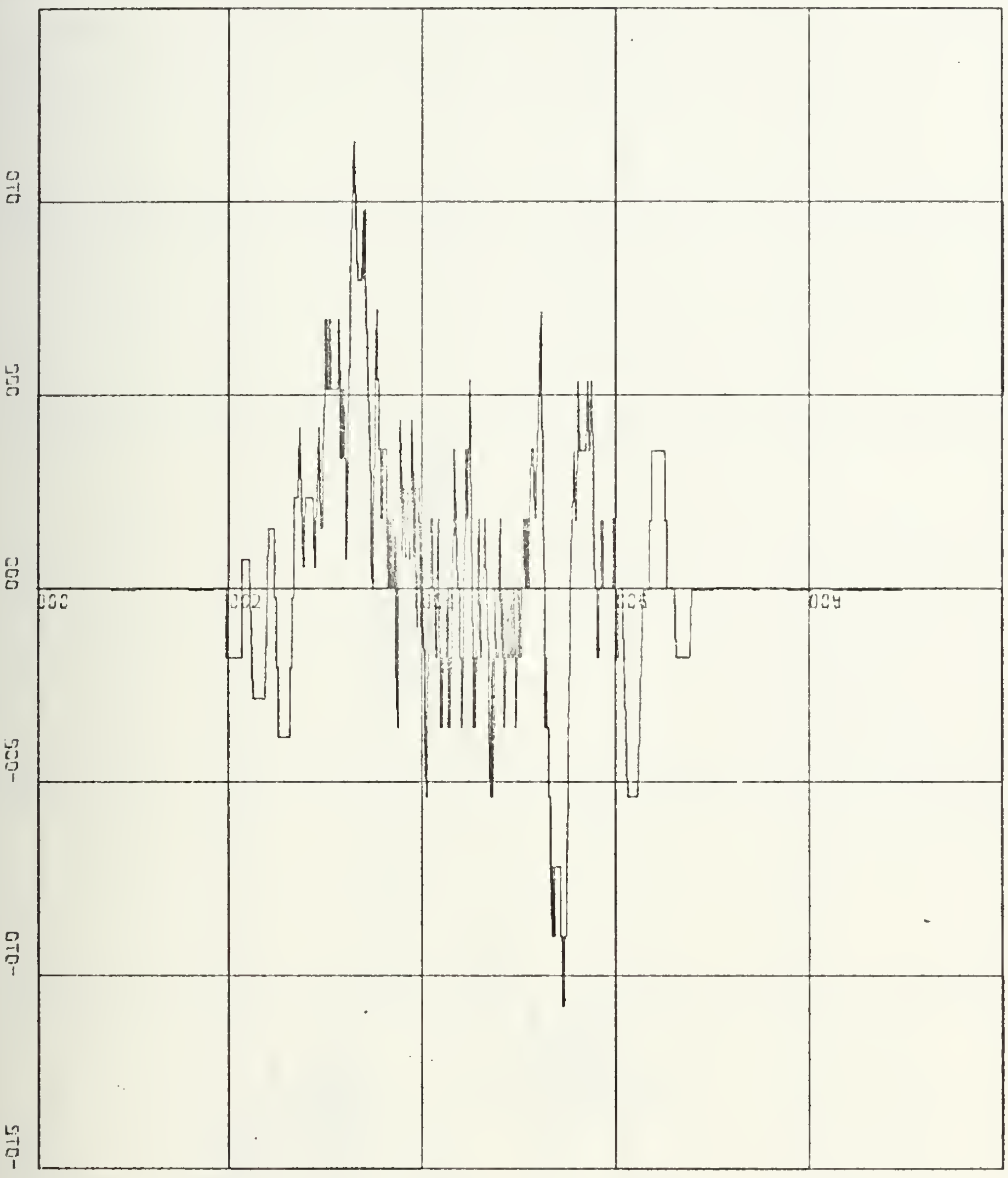
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-20

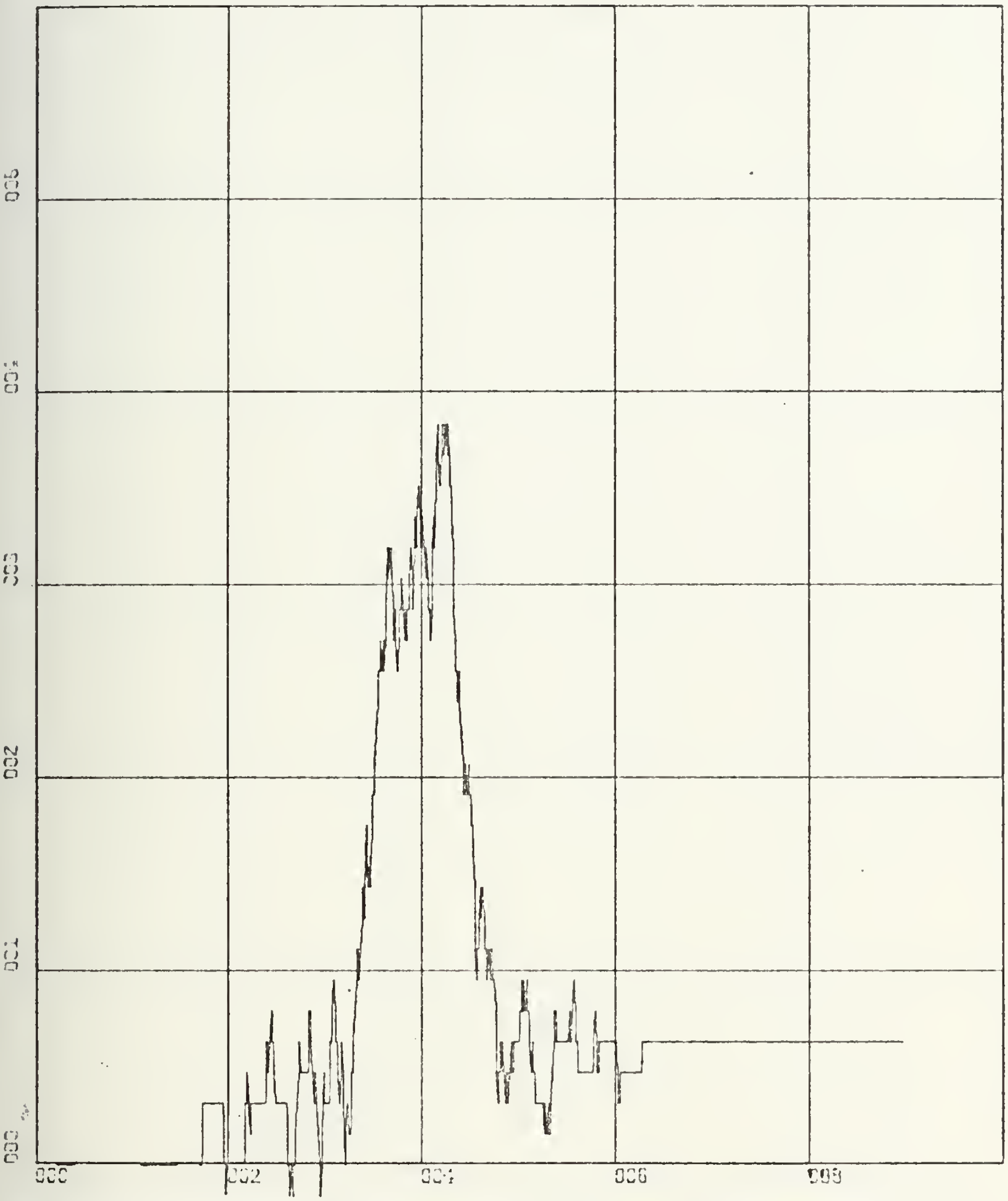
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = 10.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-21

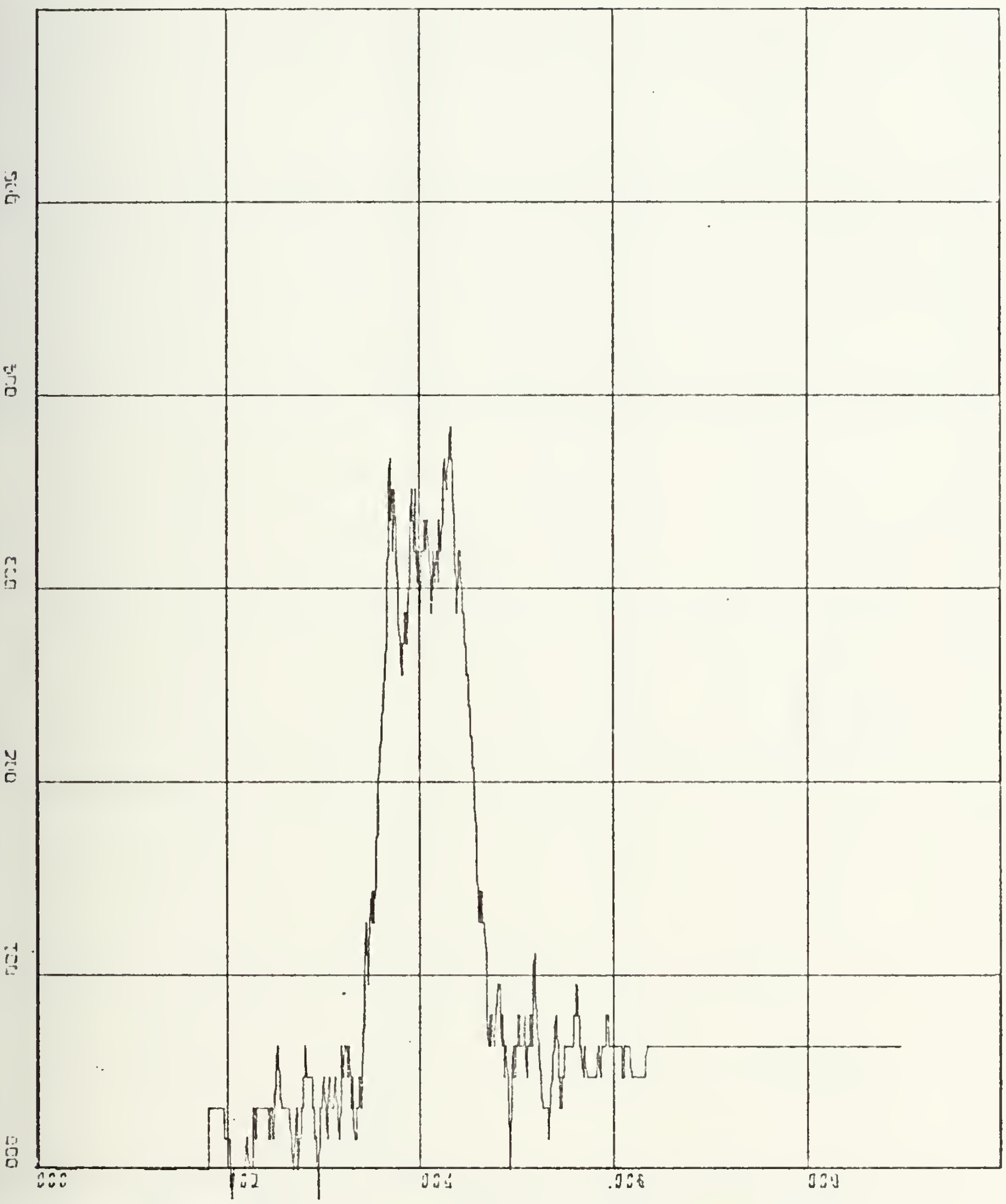
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-22

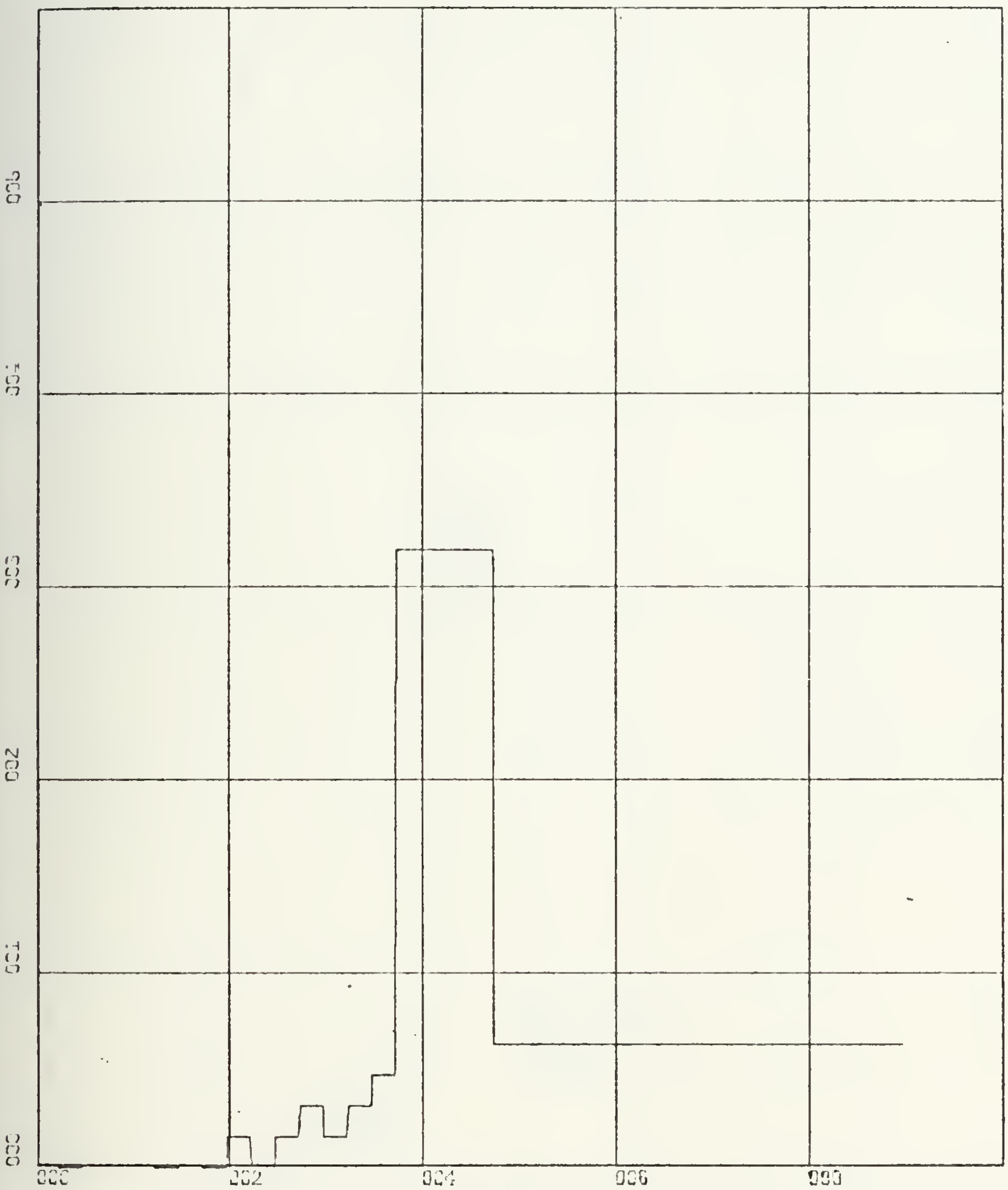
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°) ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-23

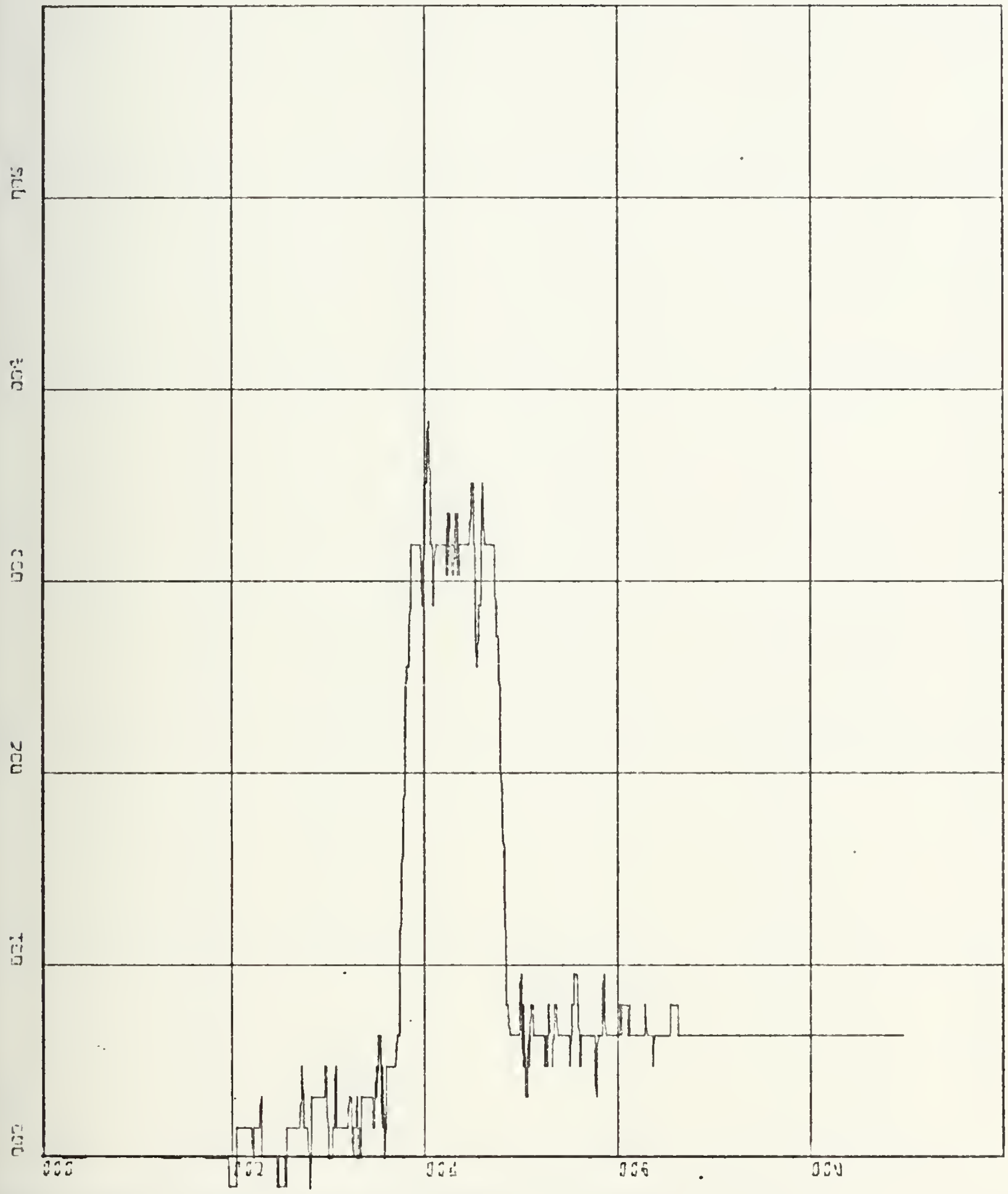
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-24

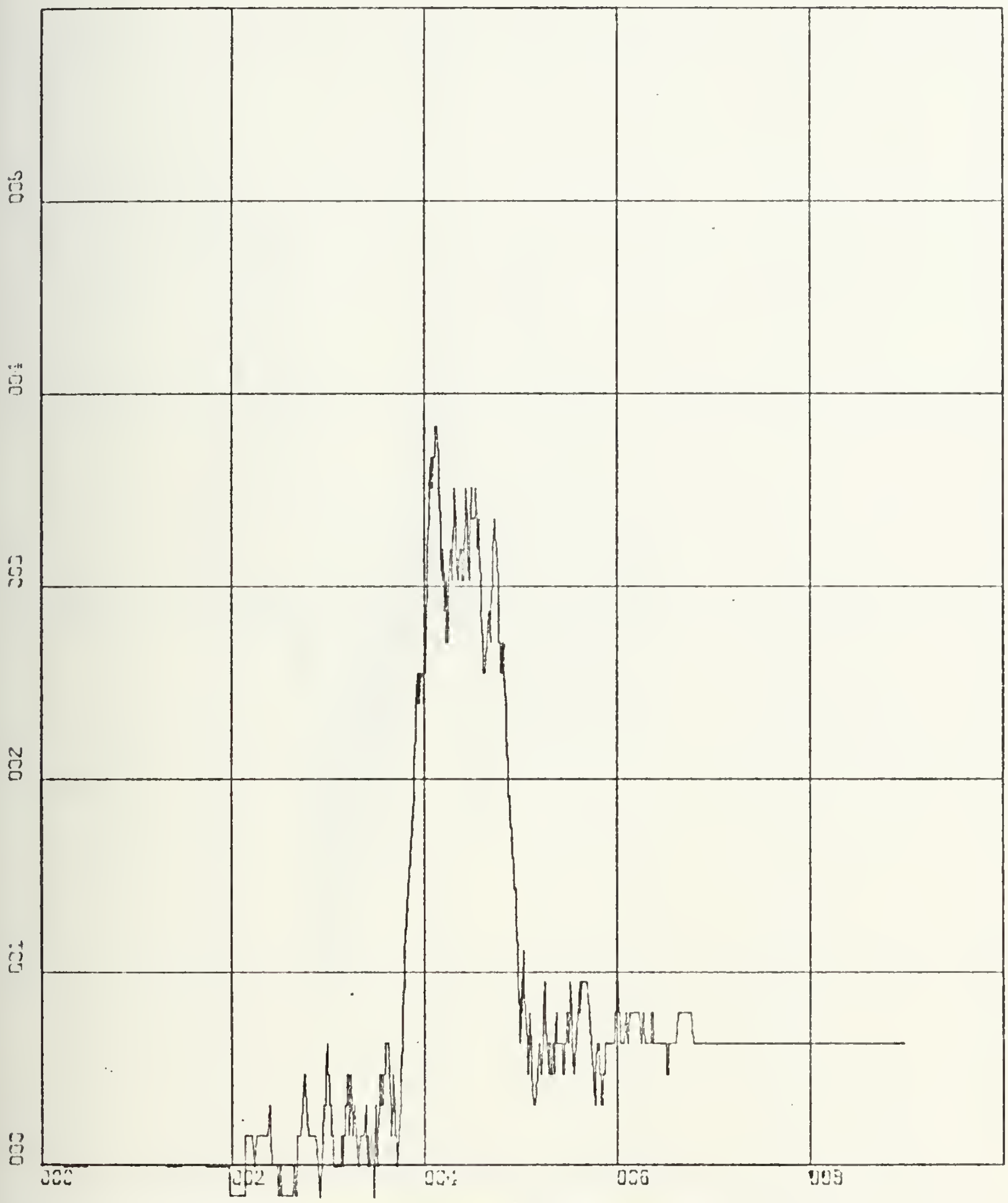
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-25

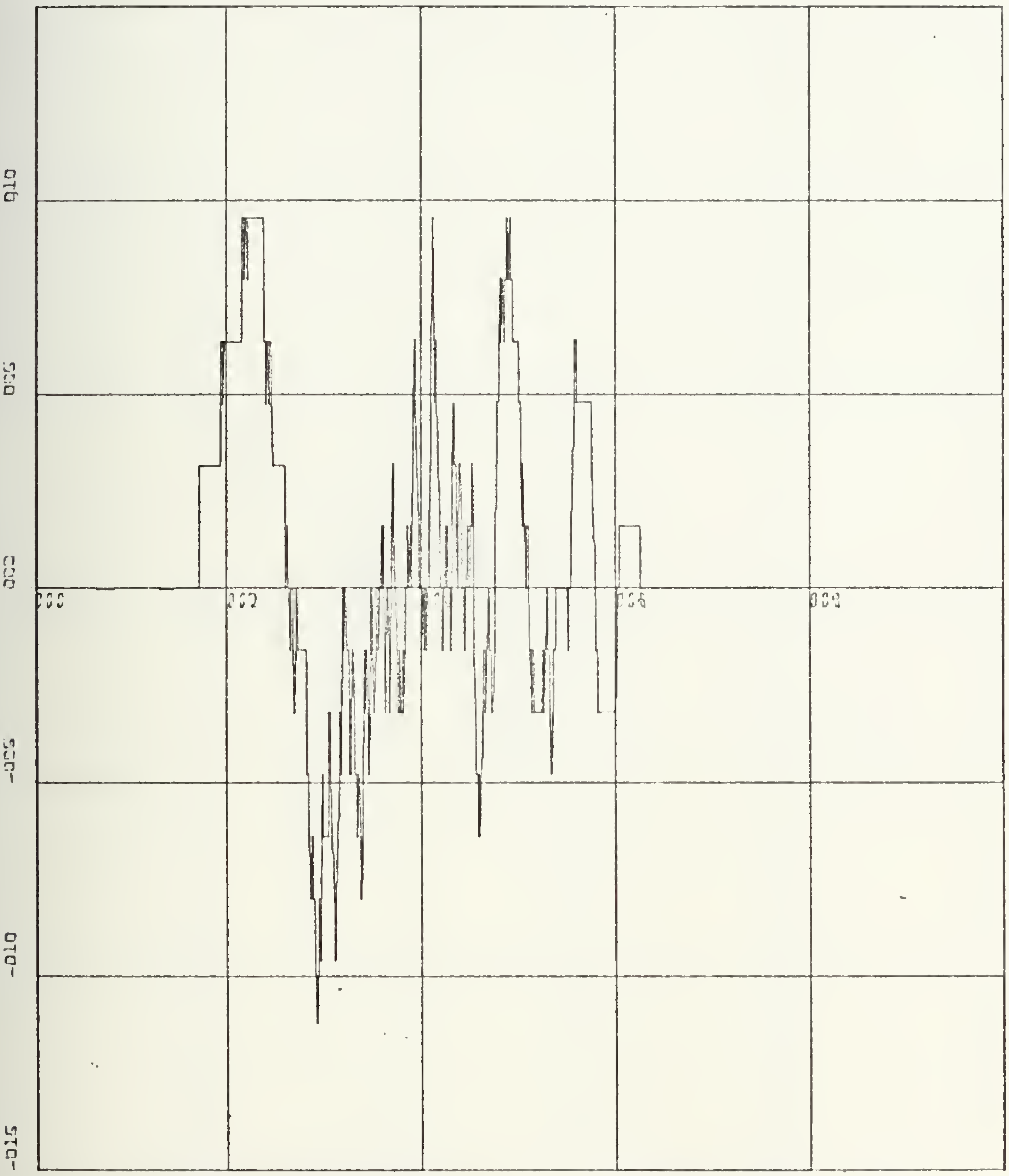
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-26

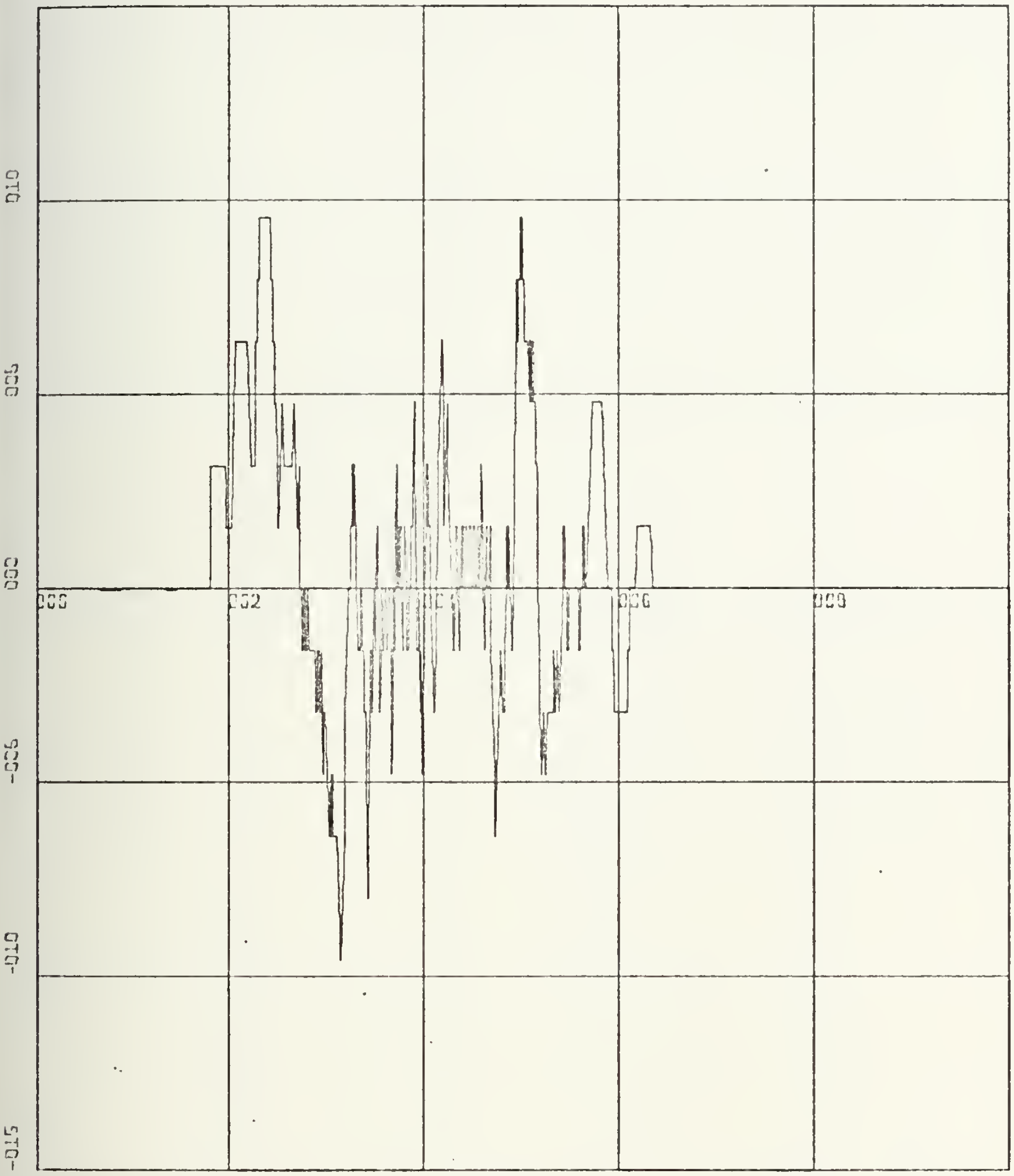
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-27

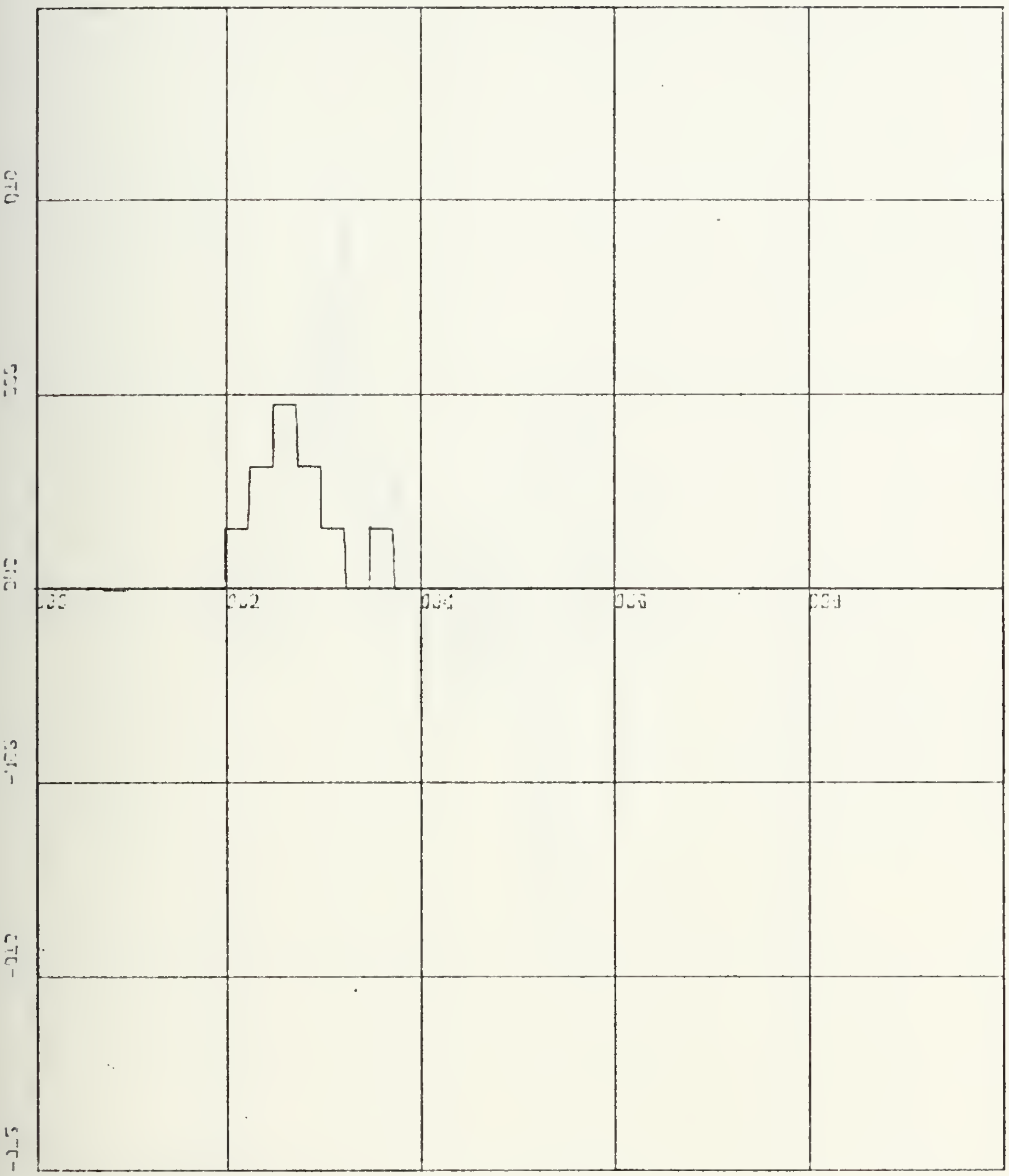
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 5 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-28

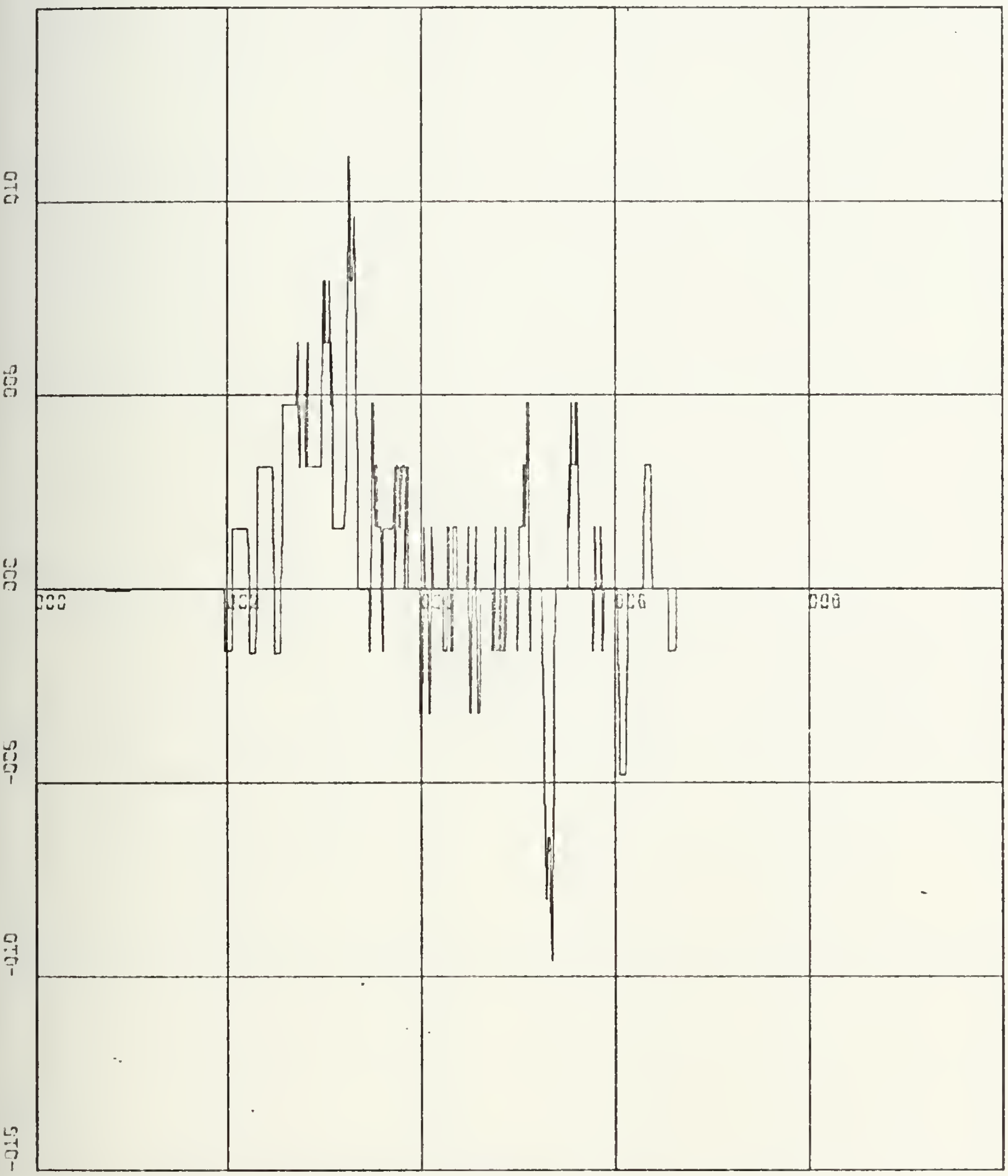
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-29

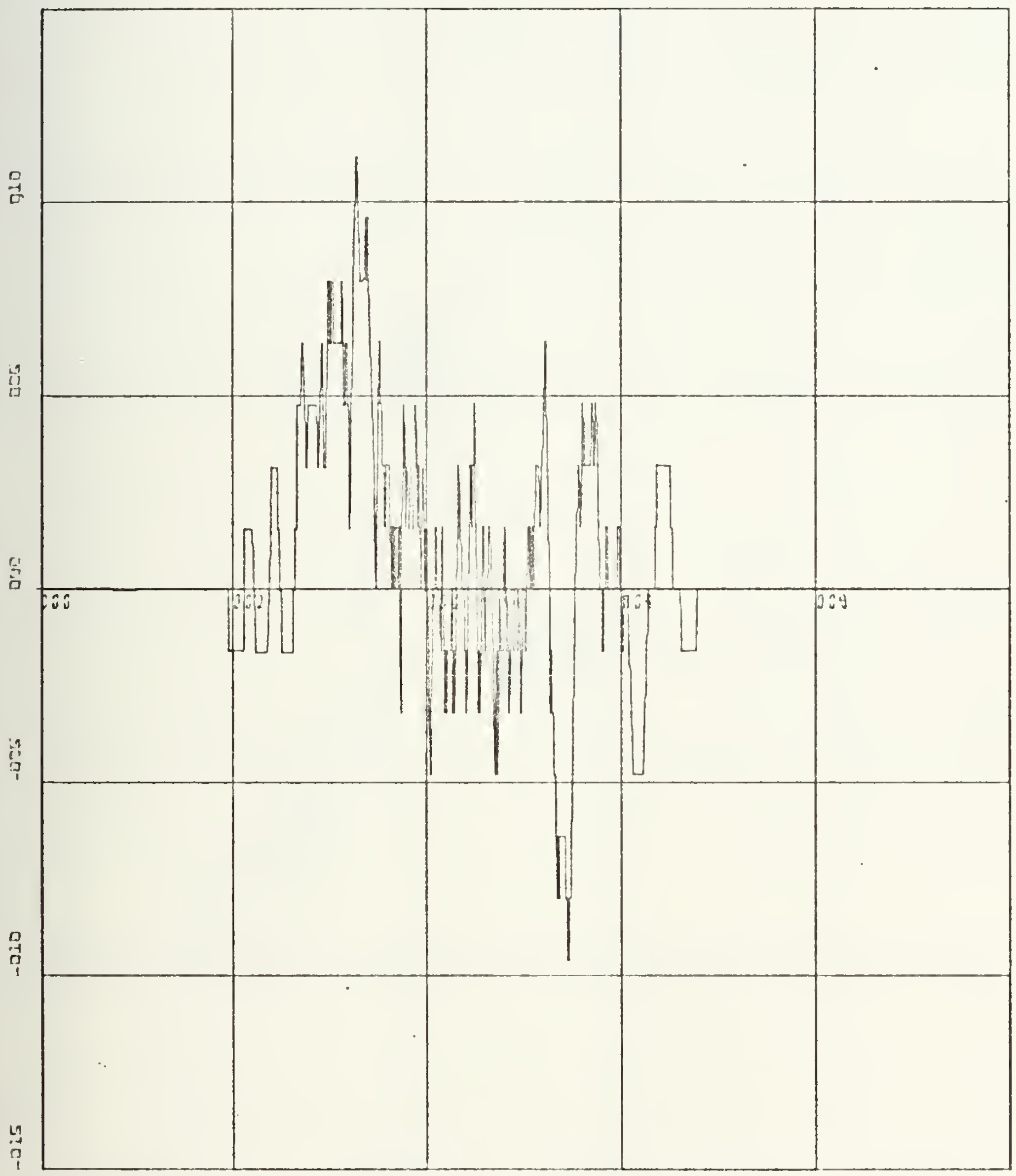
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-30

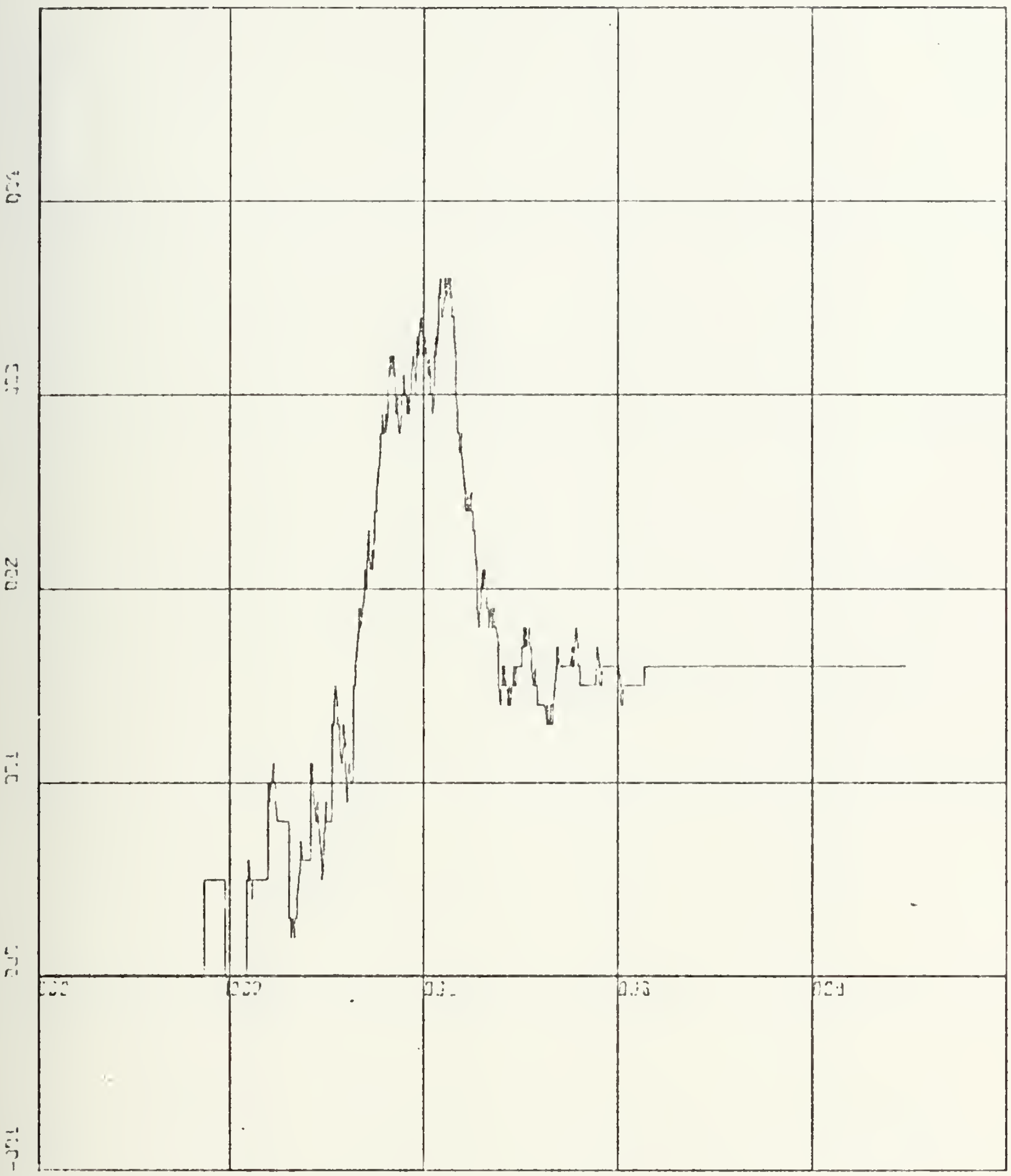
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-31

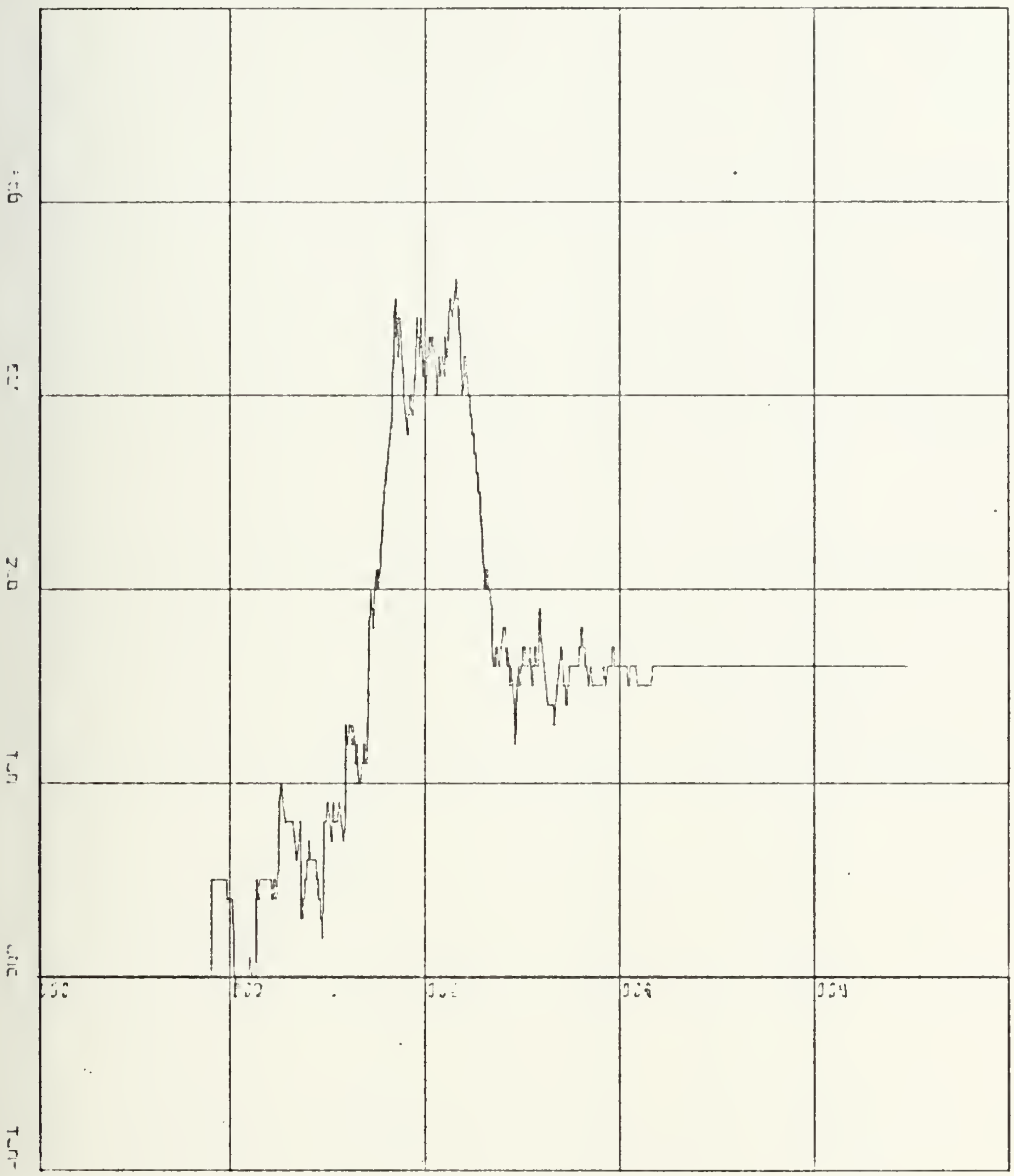
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-32

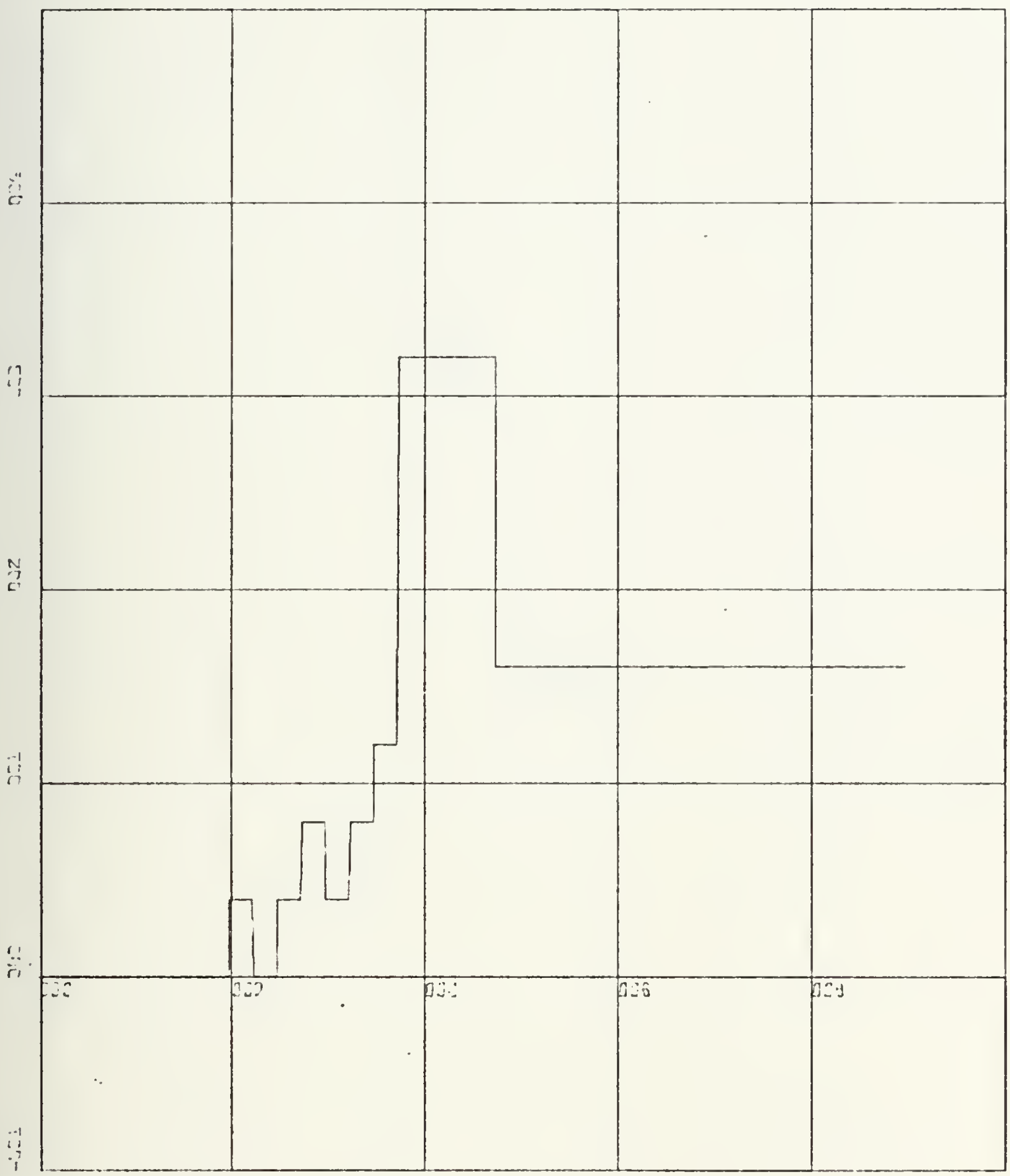
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-33

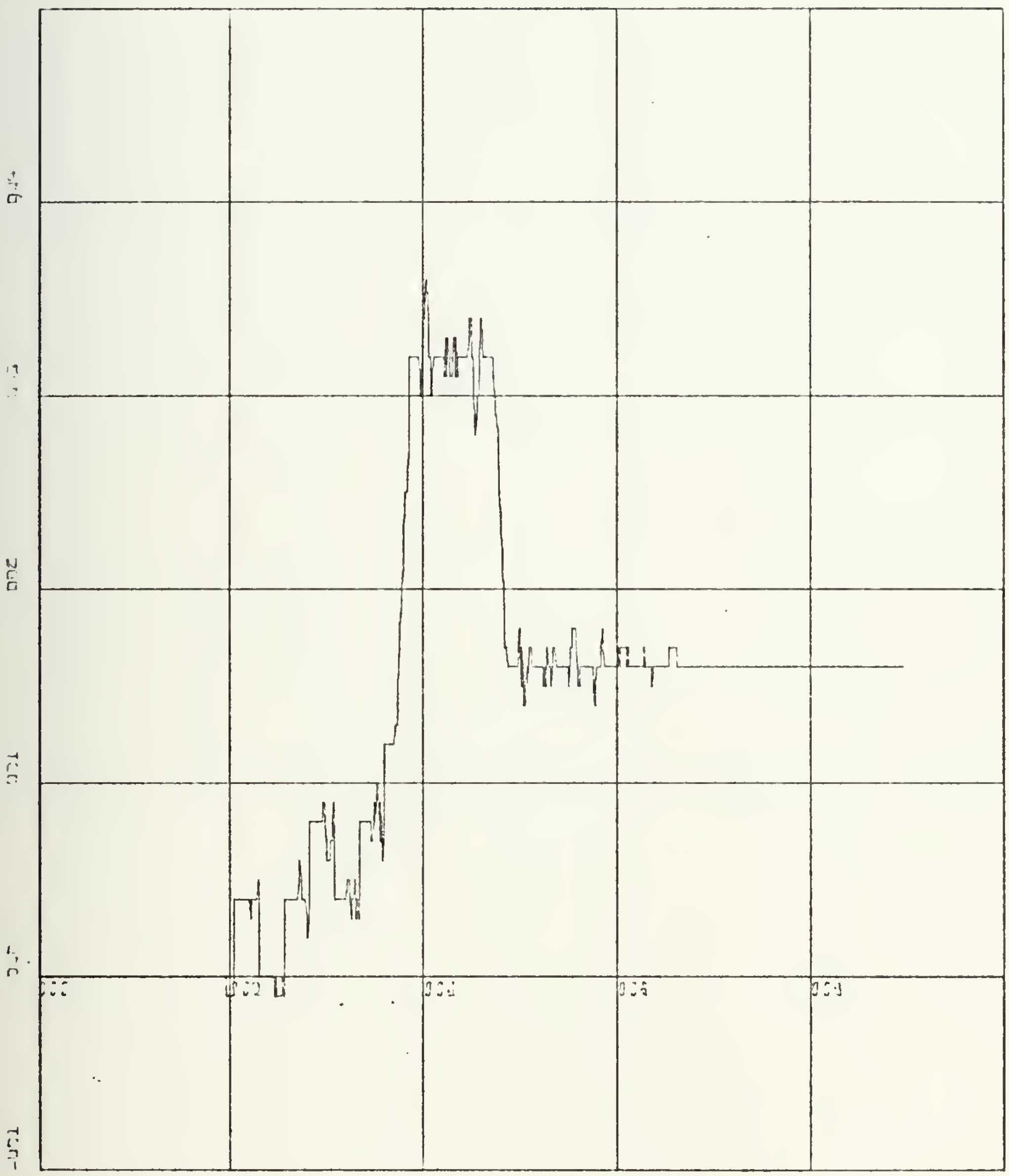
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-34

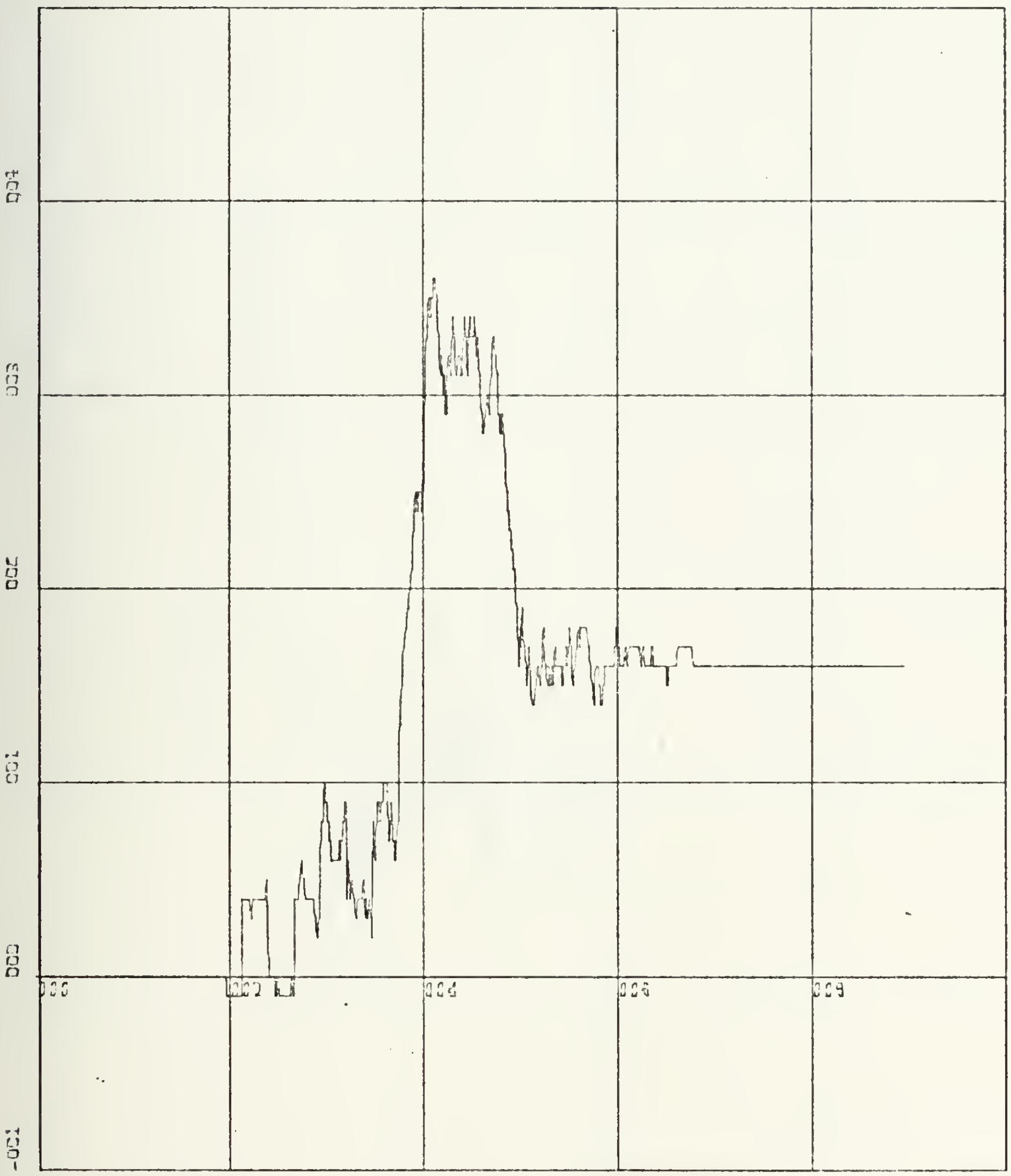
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-35

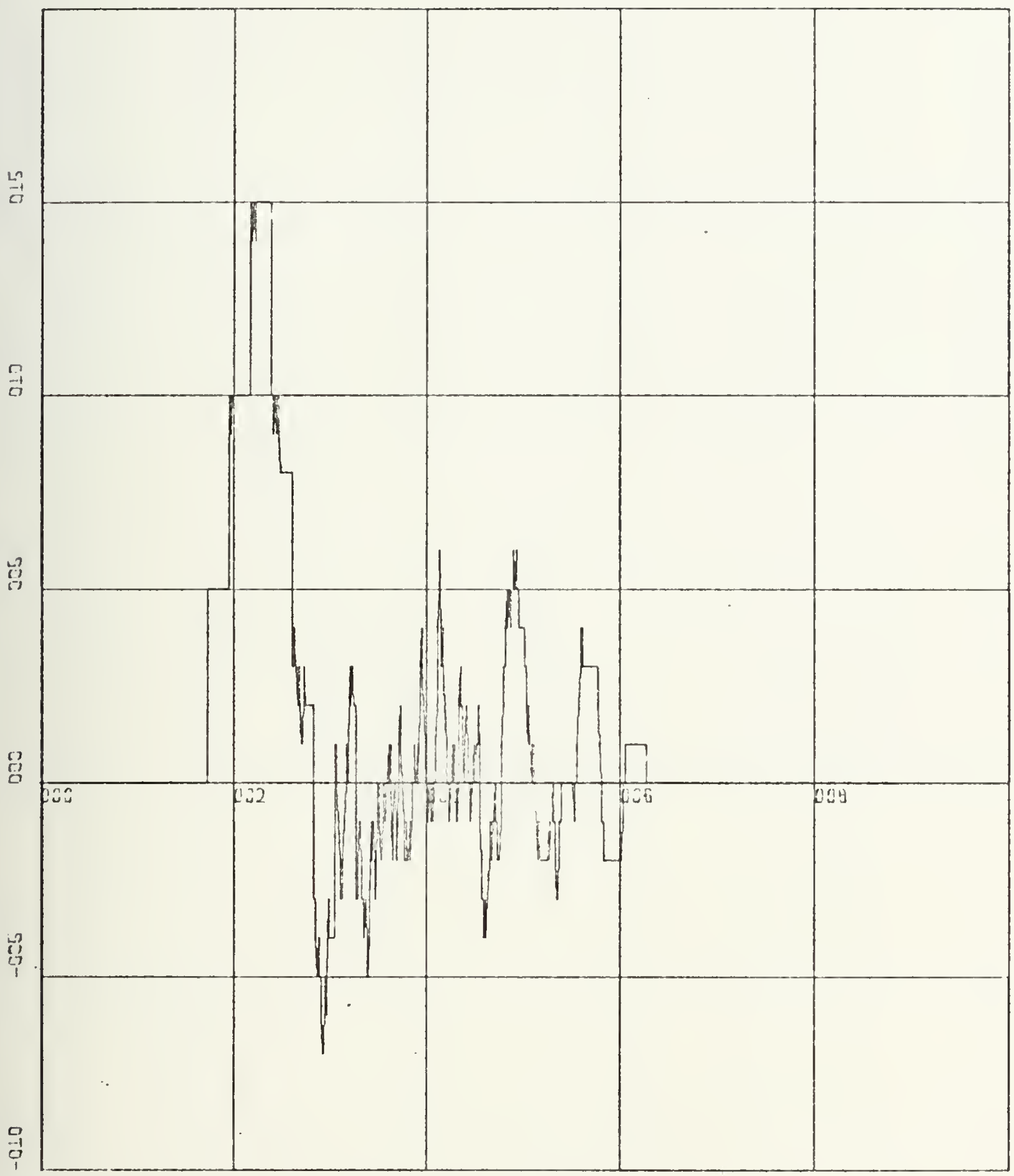
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-36

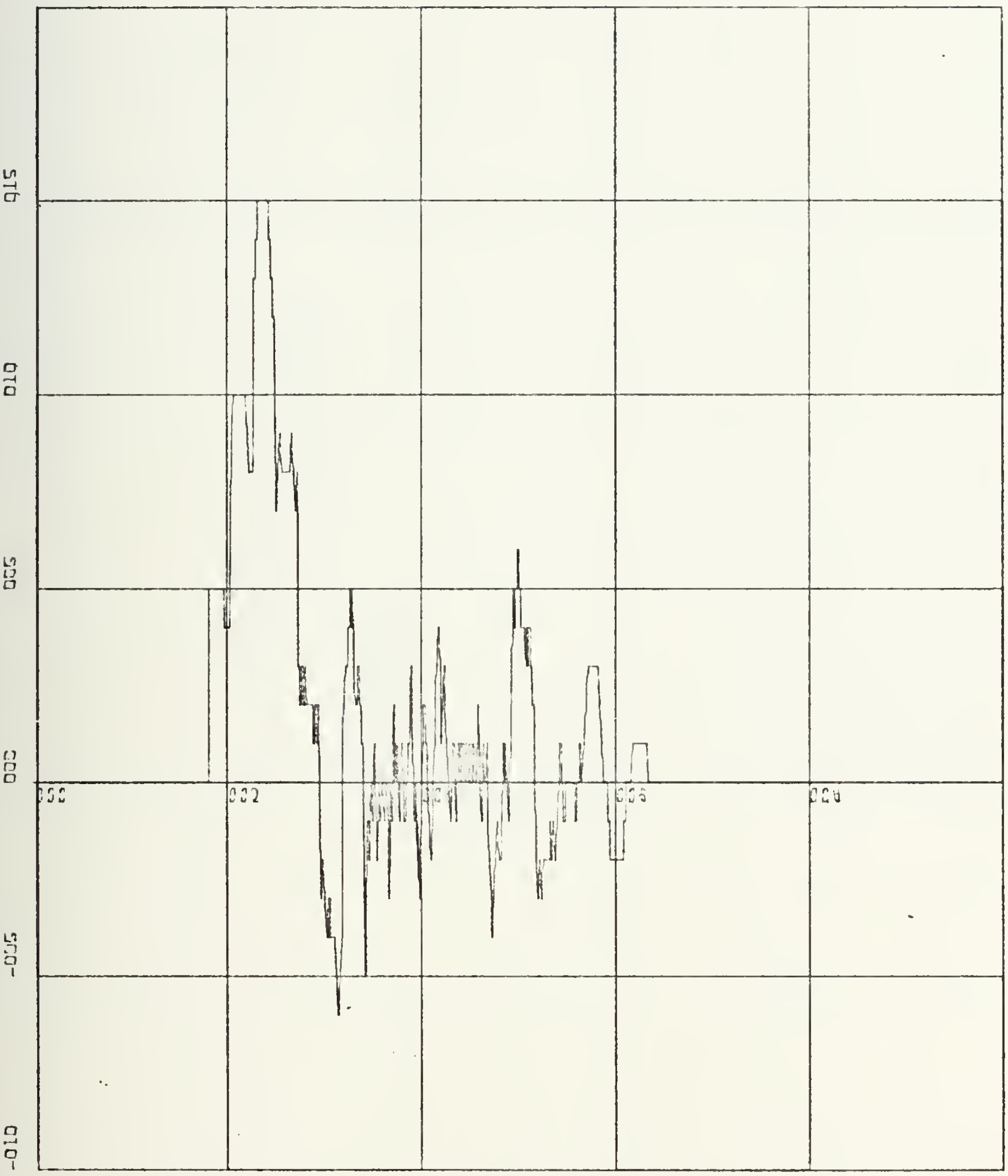
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-37

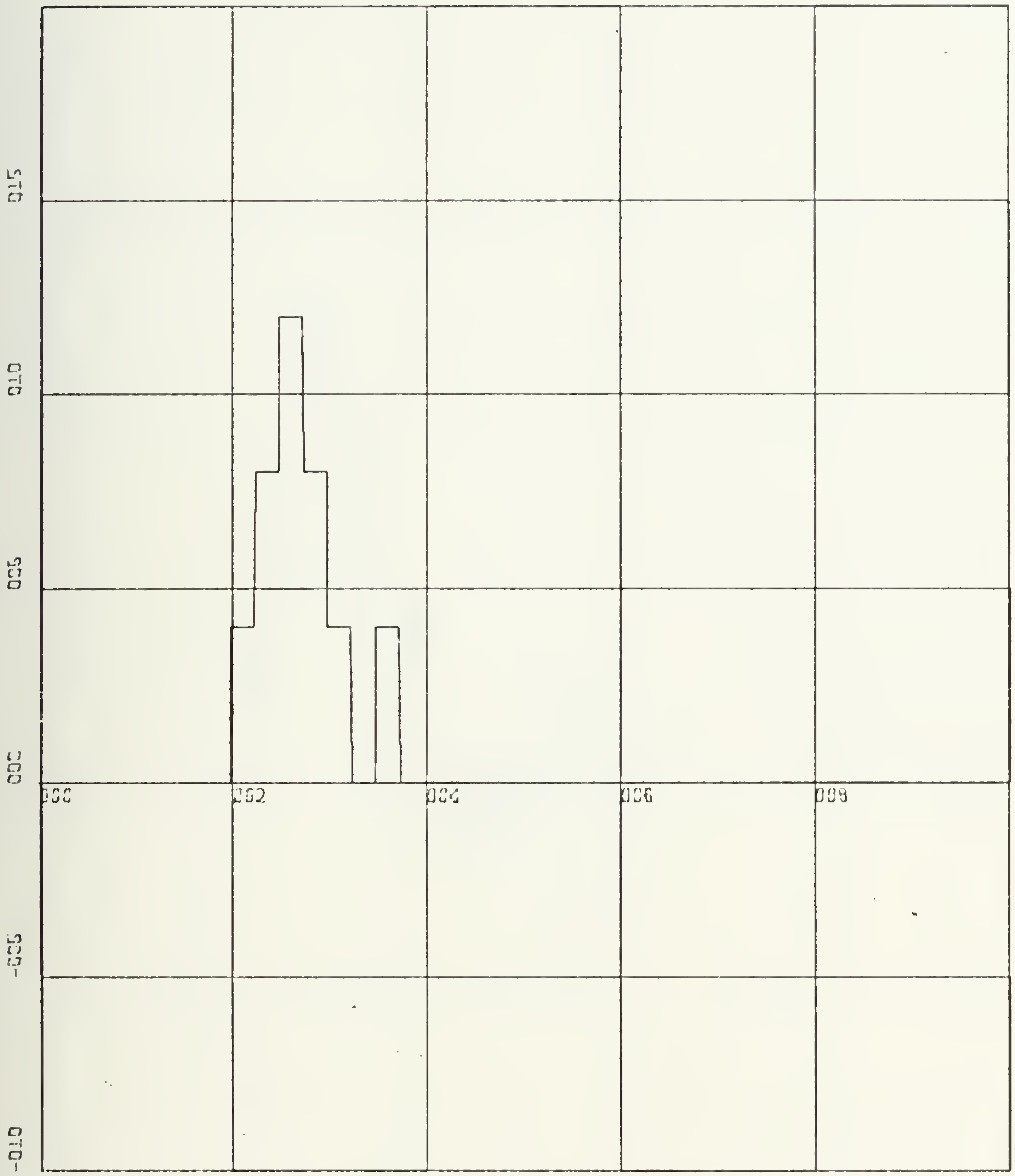
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-38

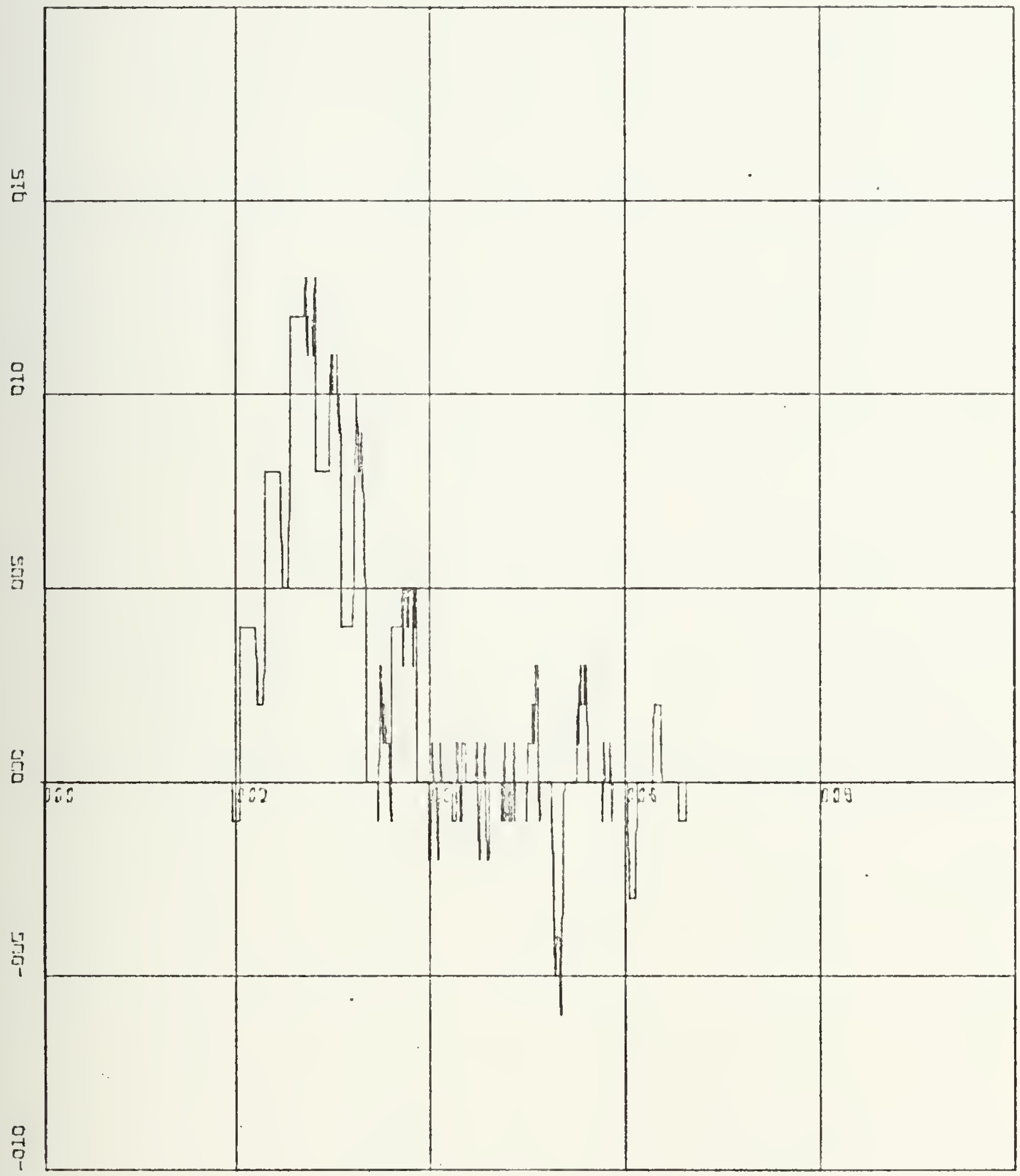
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT.
TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-39

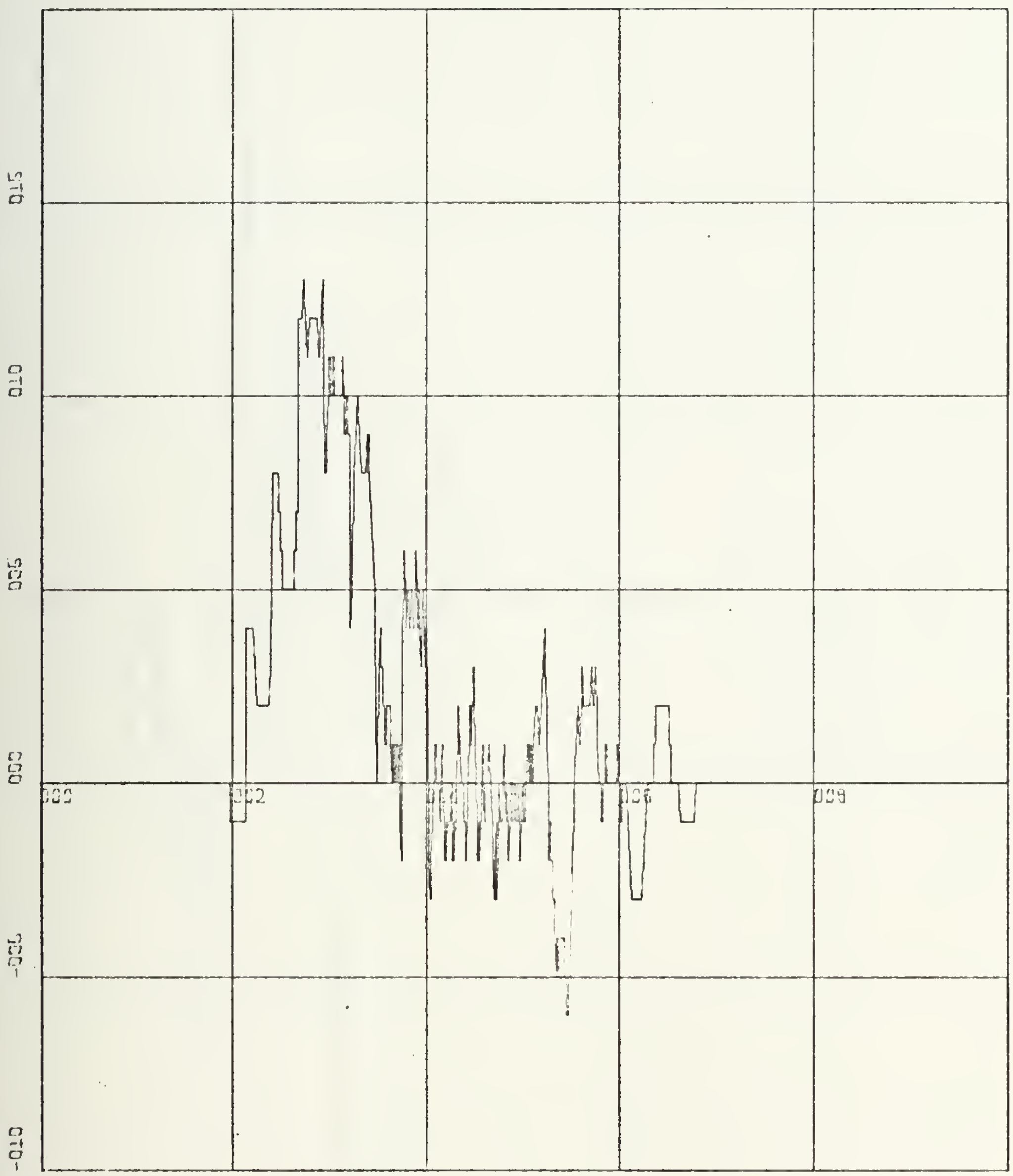
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-40

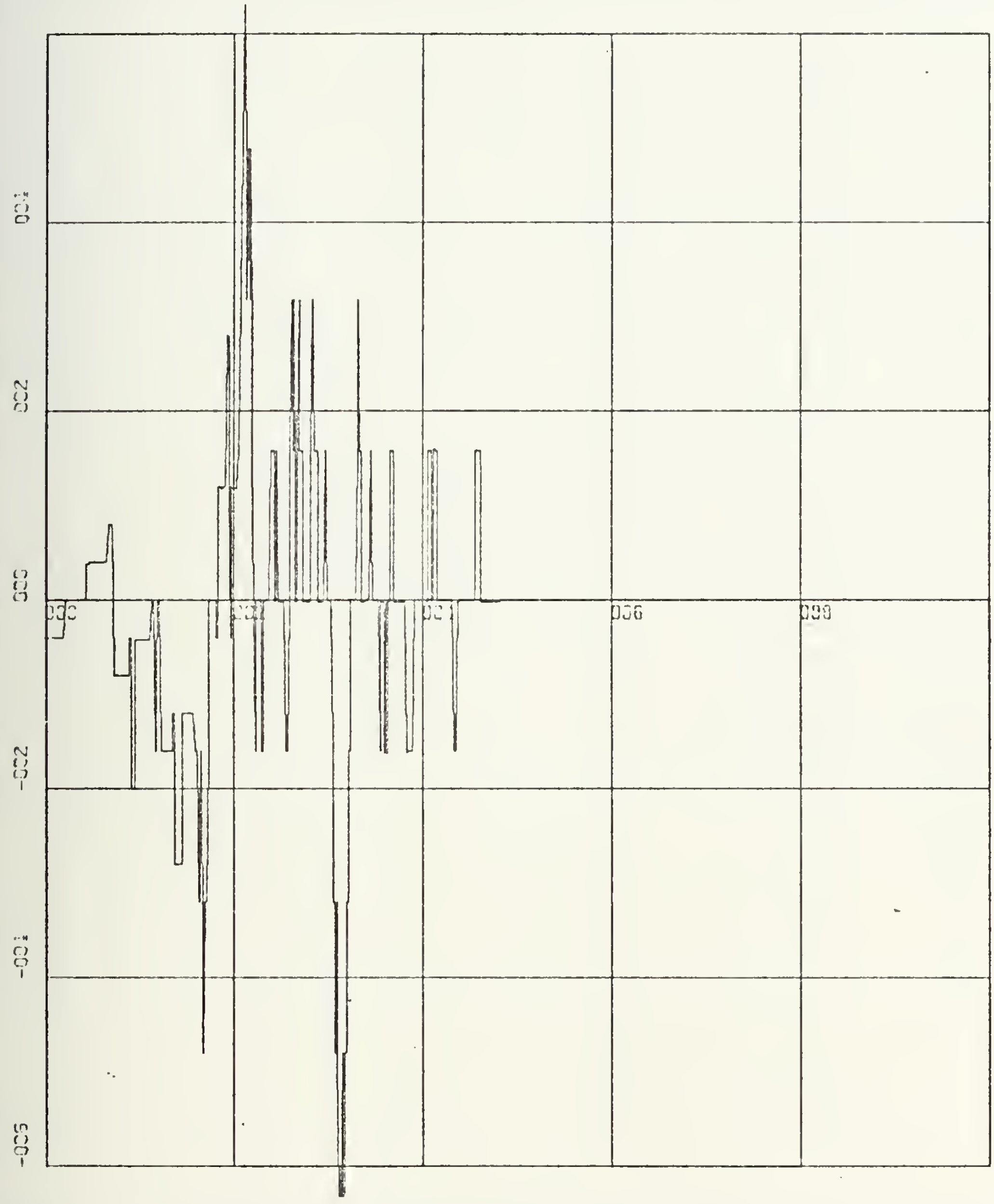
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = 2.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 100 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-41

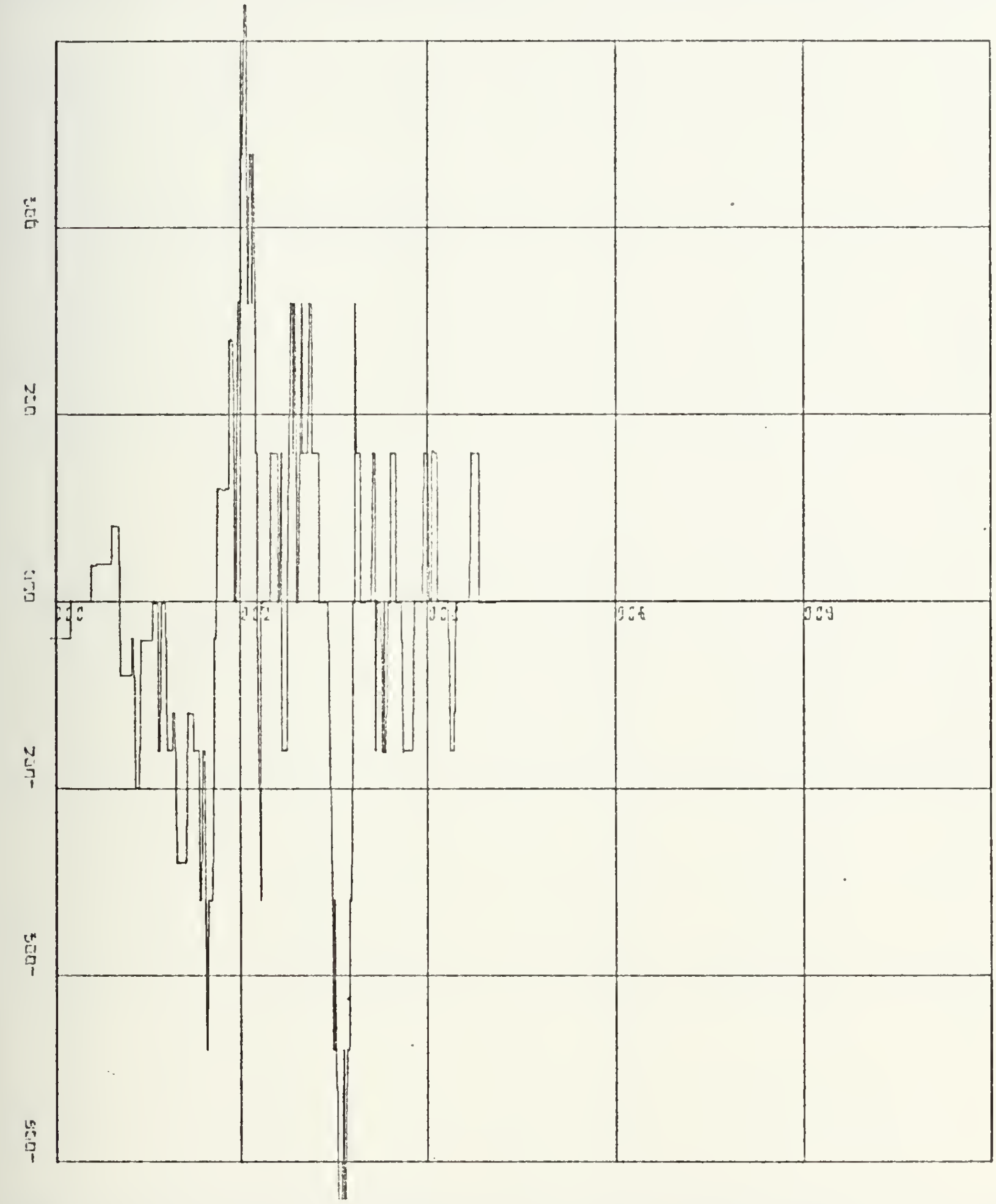
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 10, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-42

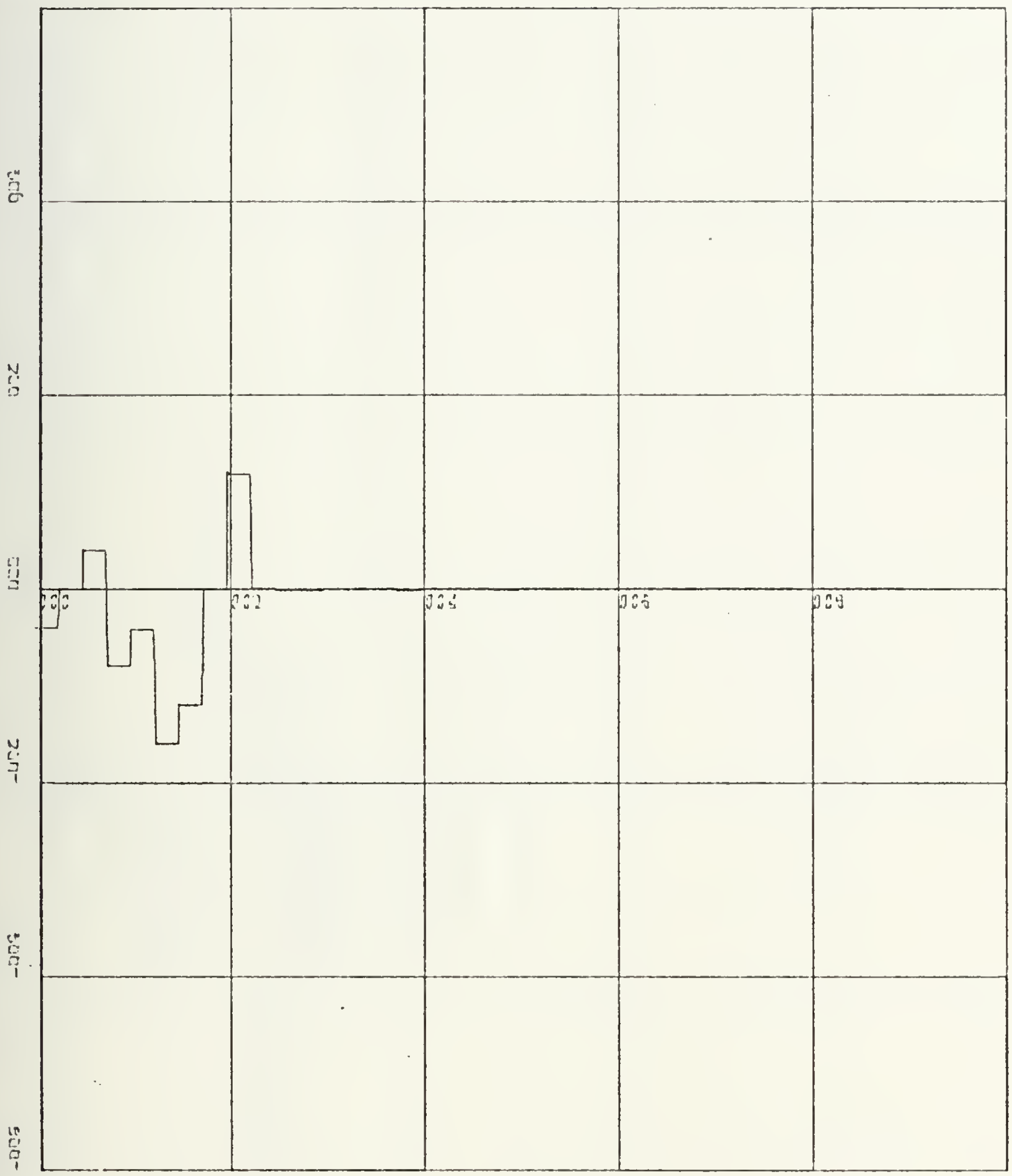
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 10, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-43

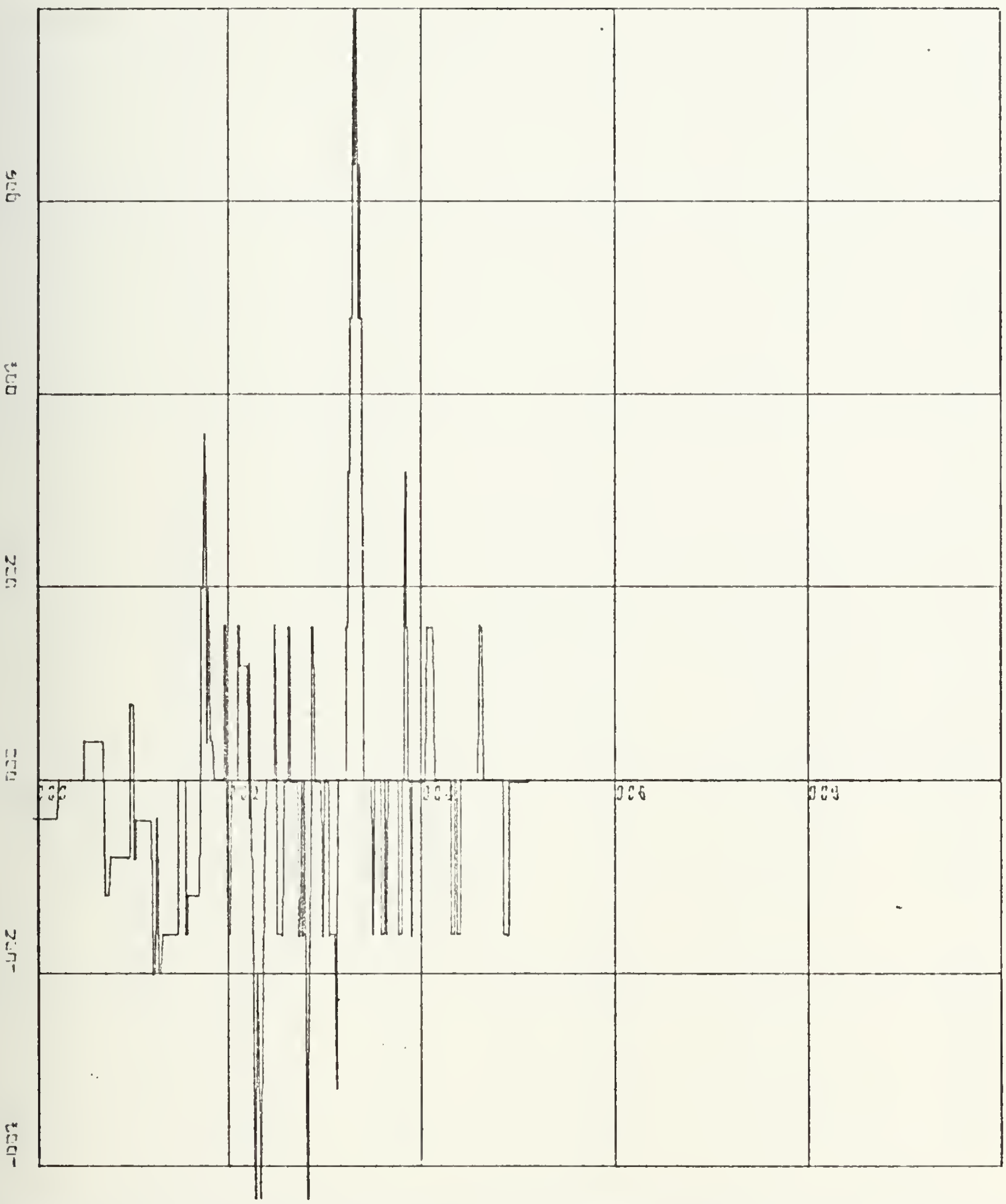
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 10, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-44

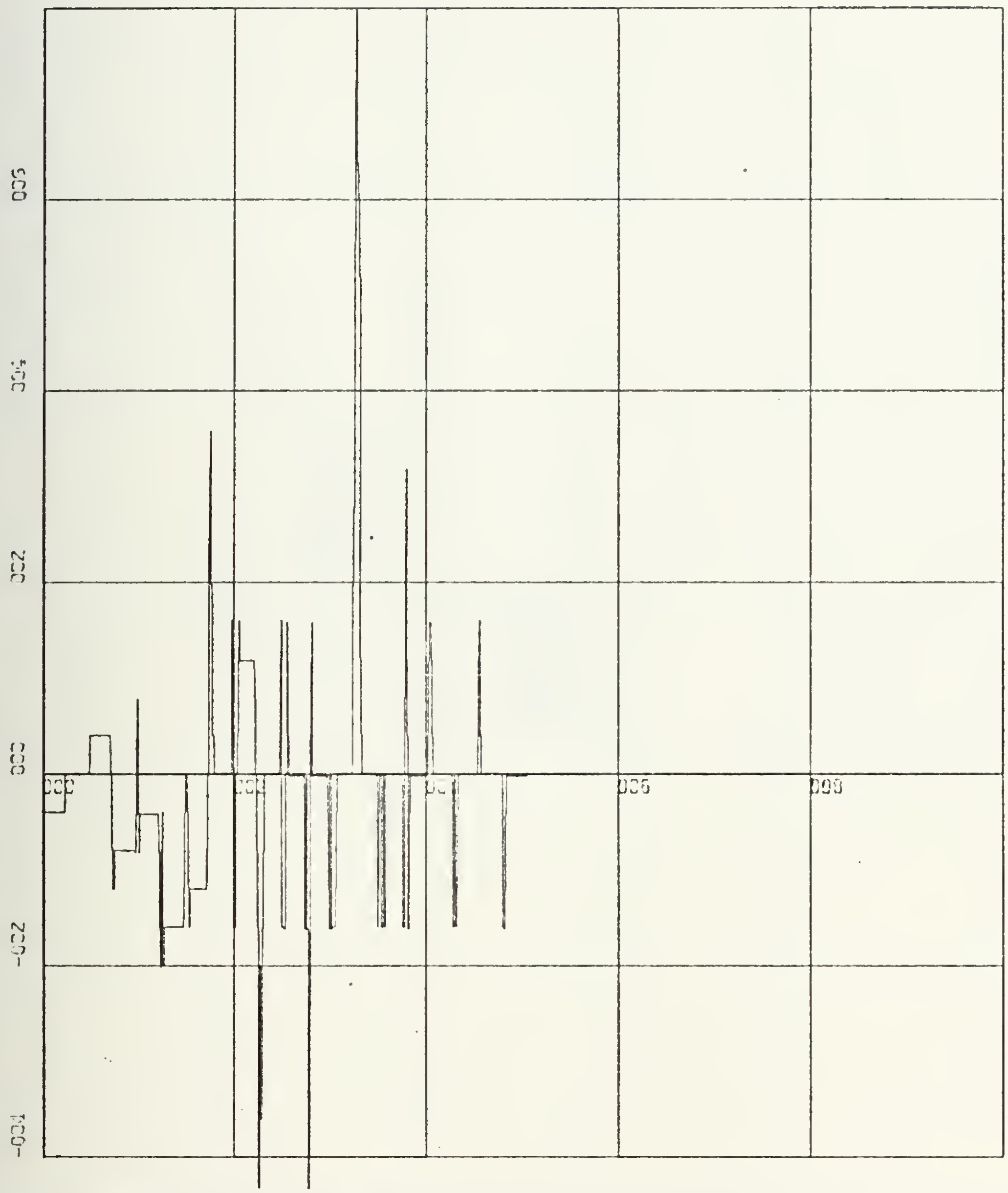
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 10, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-45

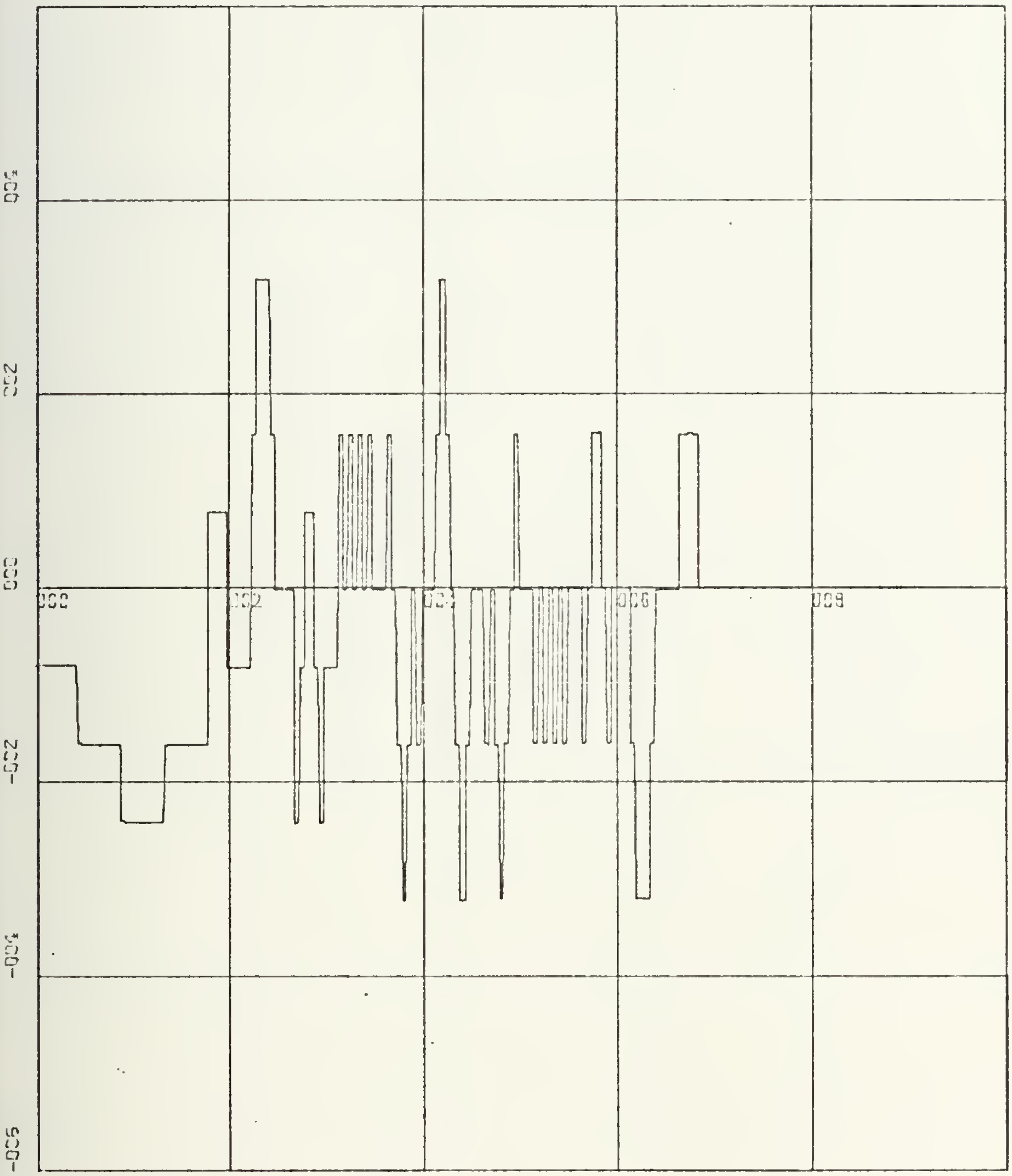
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 10, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-46

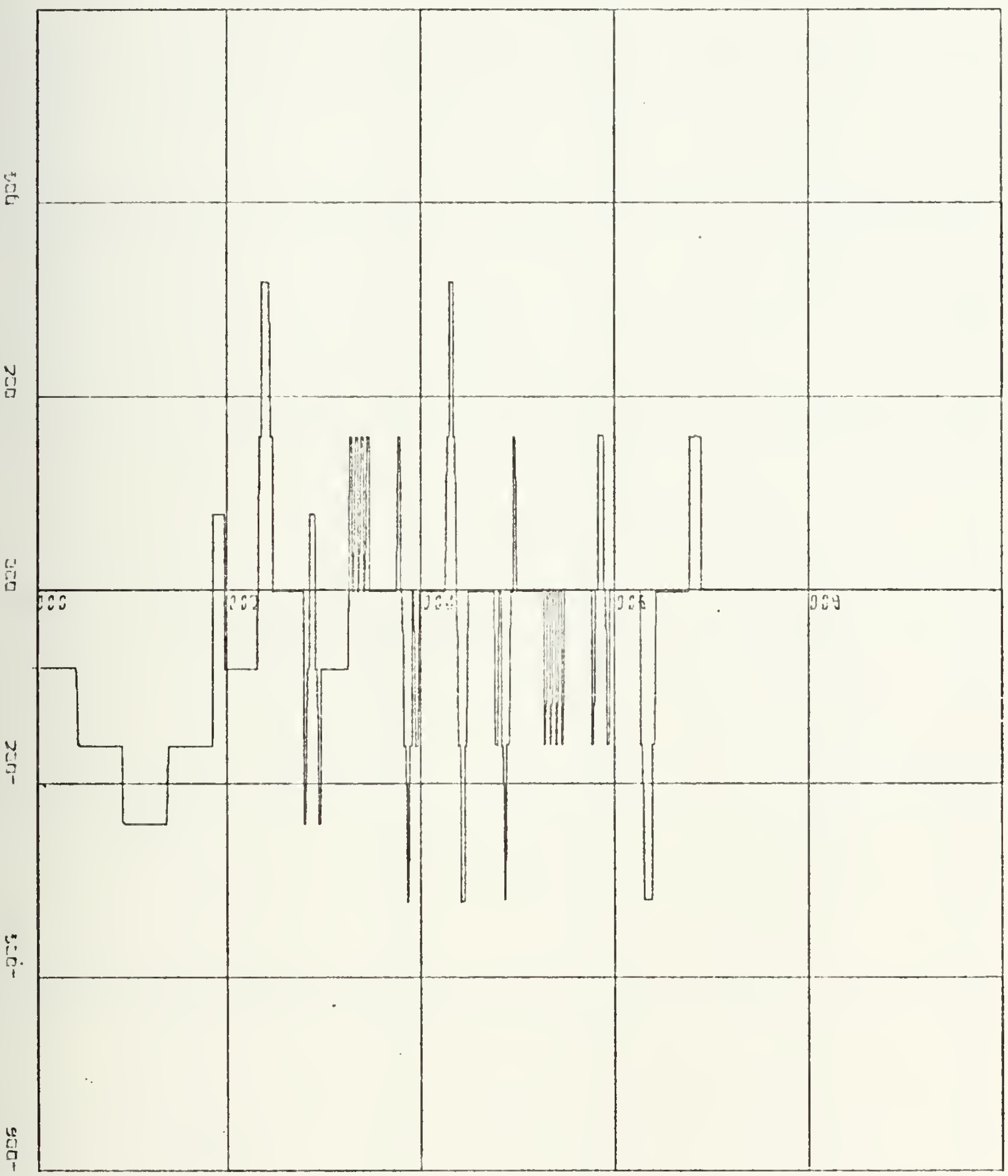
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 8, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 2 UNITS/DIV.

Figure 3-47

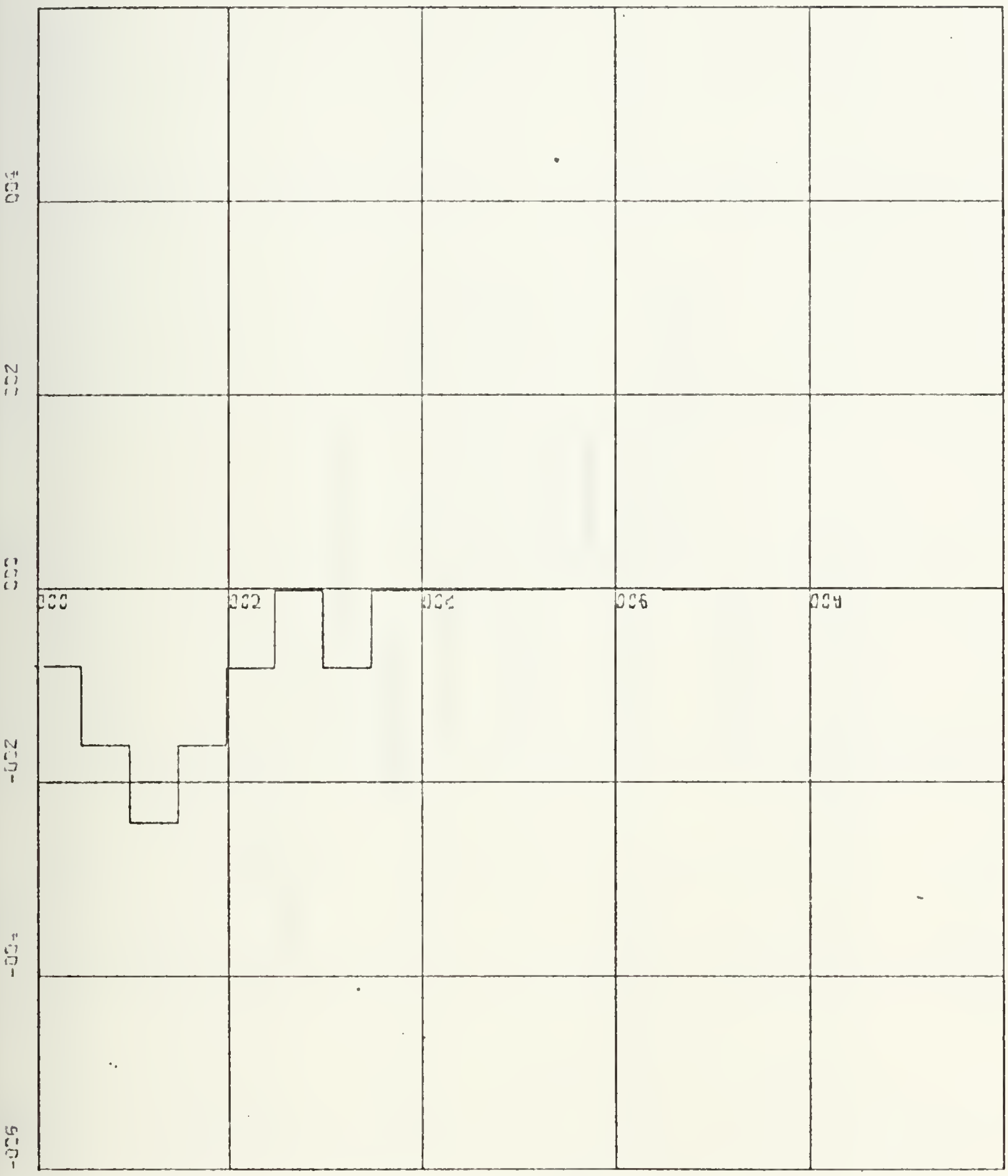
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 8, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 2 UNITS/DIV.

Figure 3-48

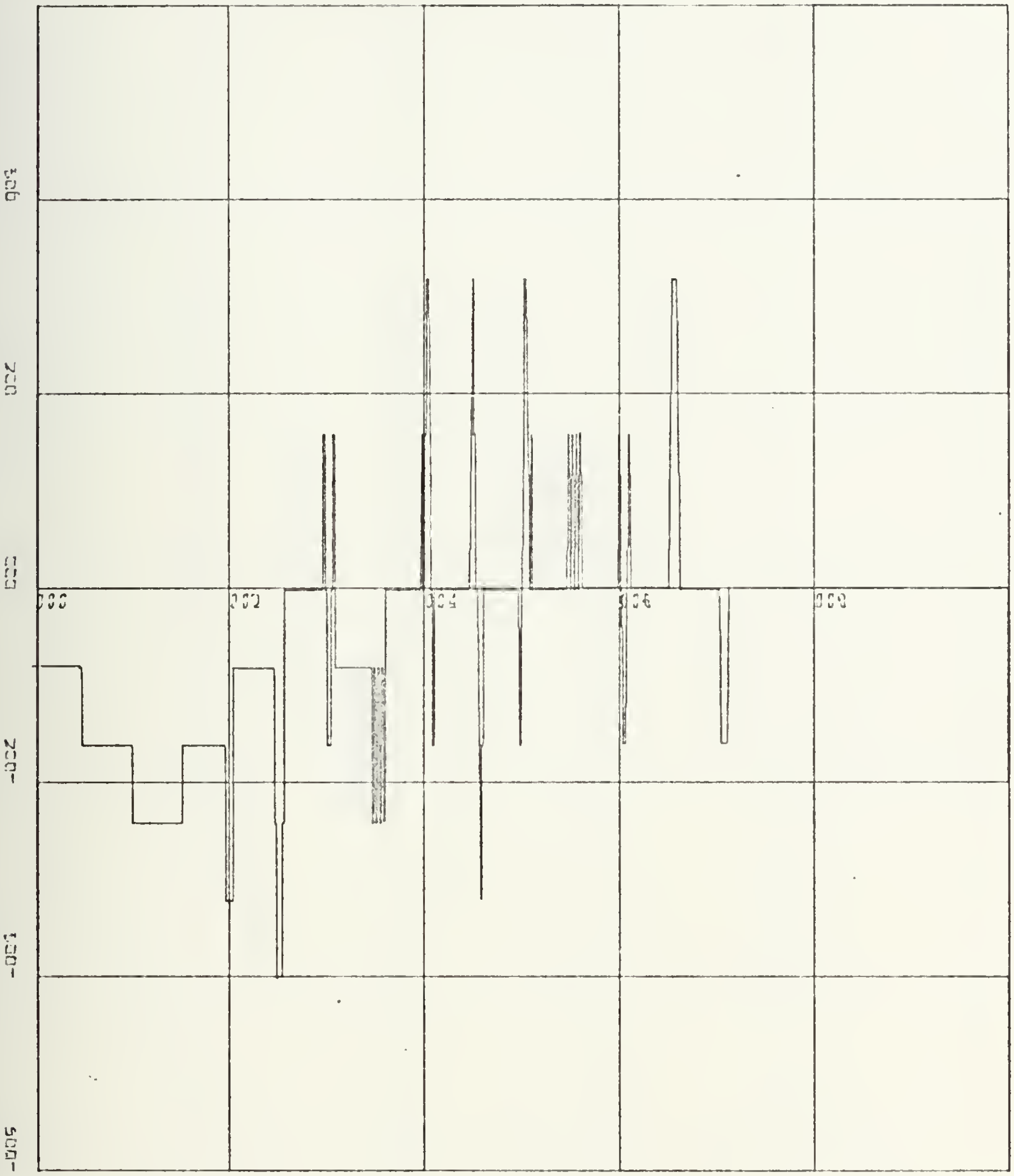
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 8, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 2 UNITS/DIV.

Figure 3-49

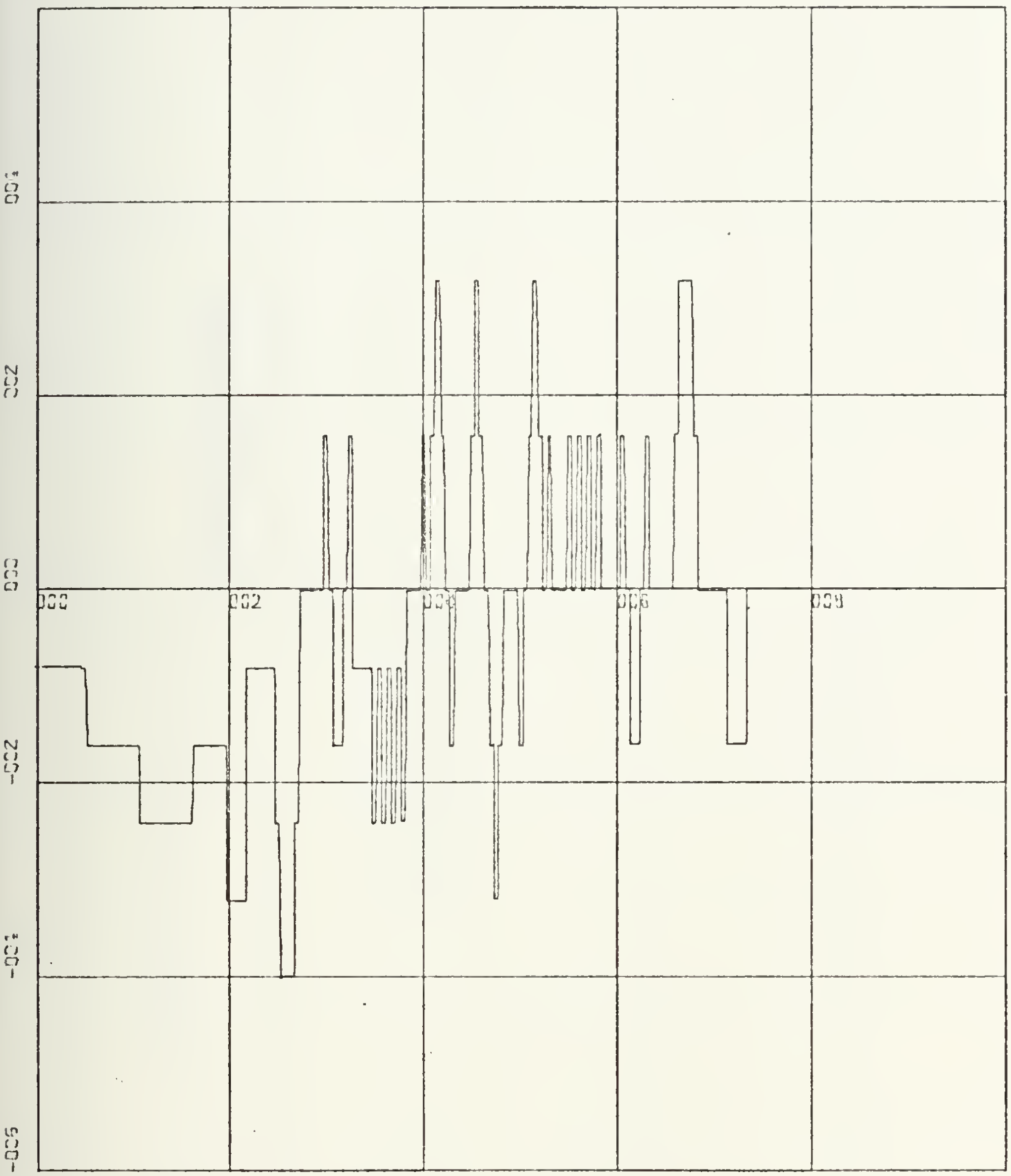
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 8, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 2 UNITS/DIV.

Figure 3-50

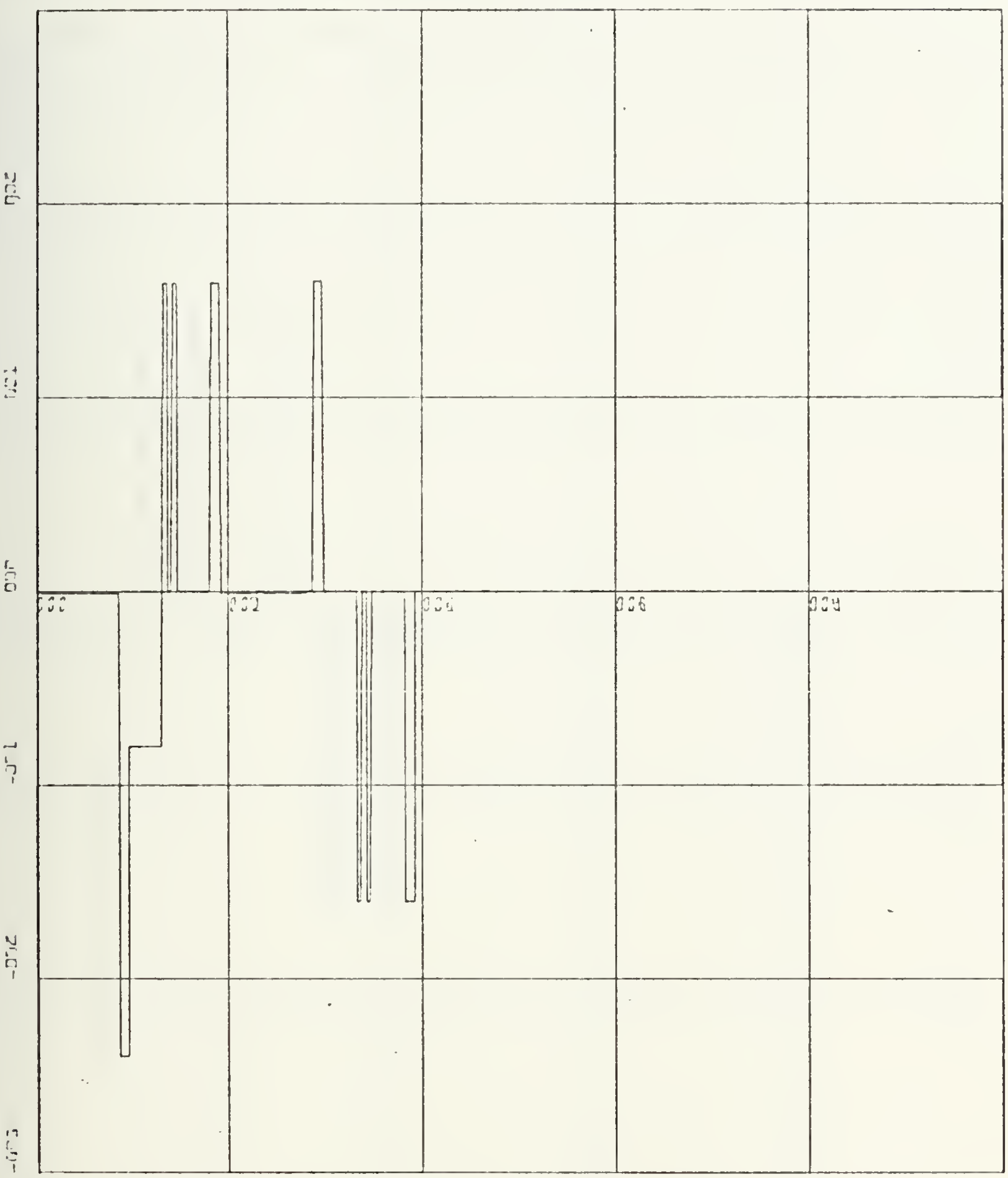
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 8, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 2 UNITS/DIV.

Figure 3-51

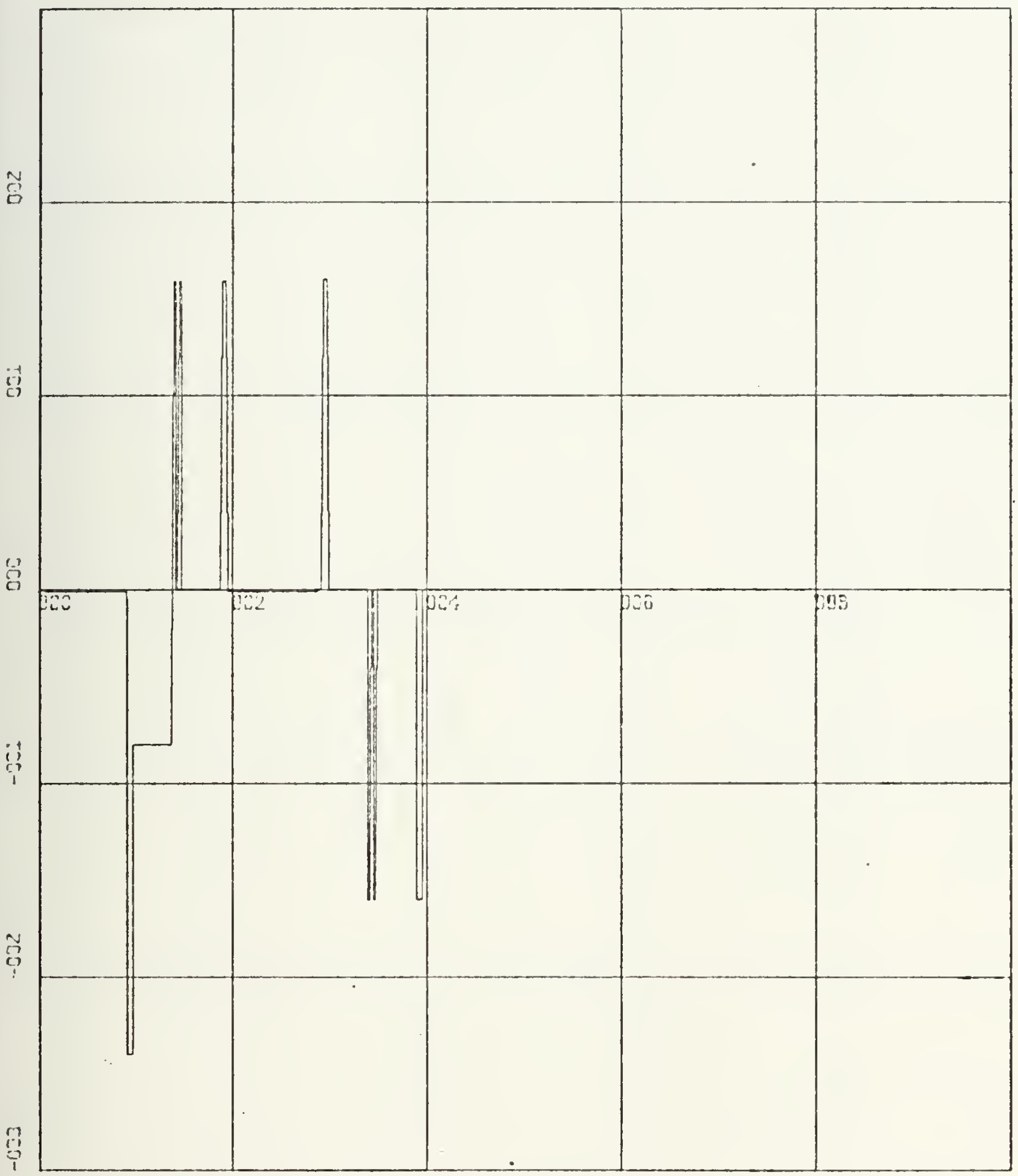
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 4, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH ---- 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 1 UNIT/DIV.

Figure 3-52

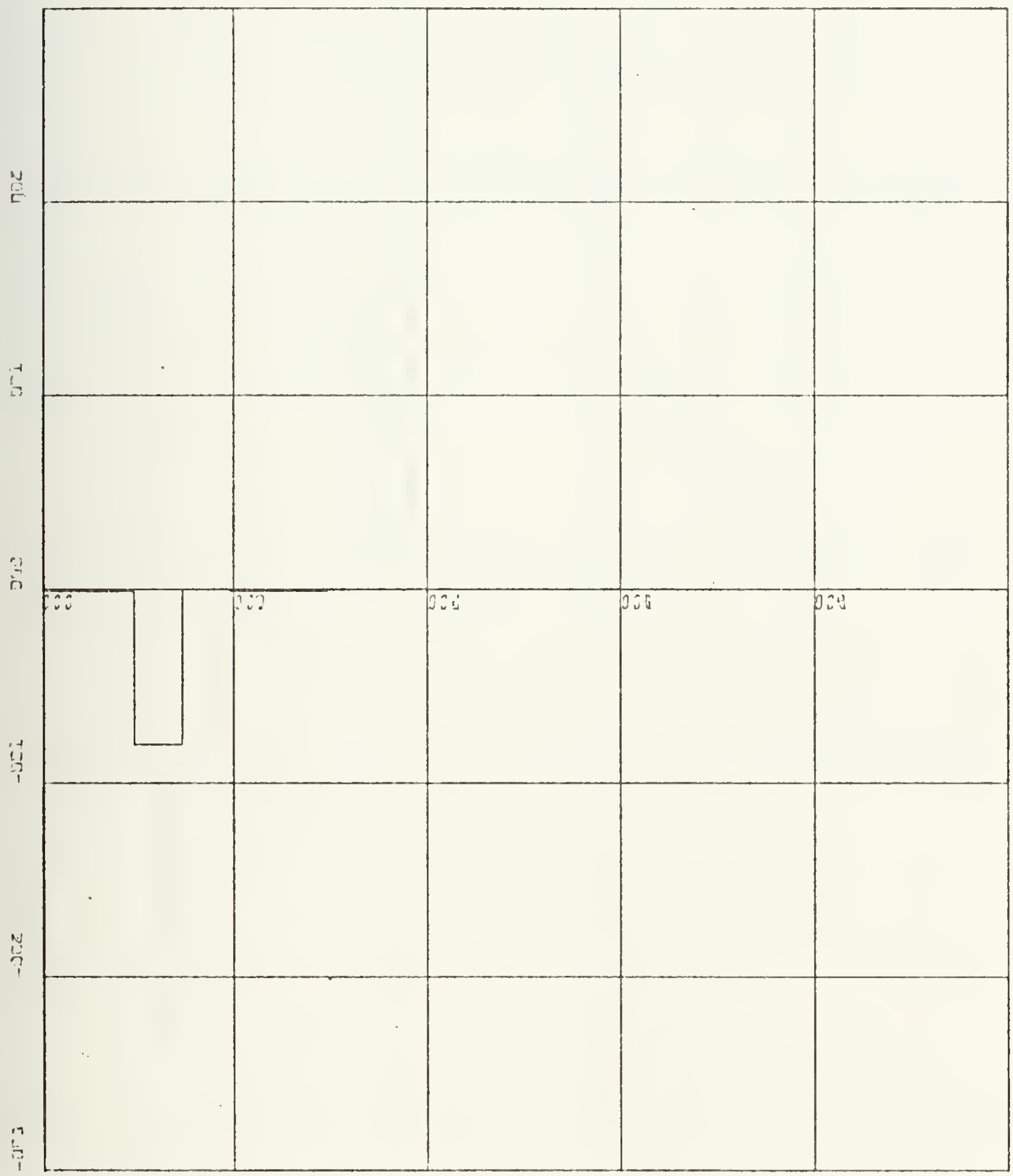
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 4, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 1 UNIT/DIV.

Figure 3-53

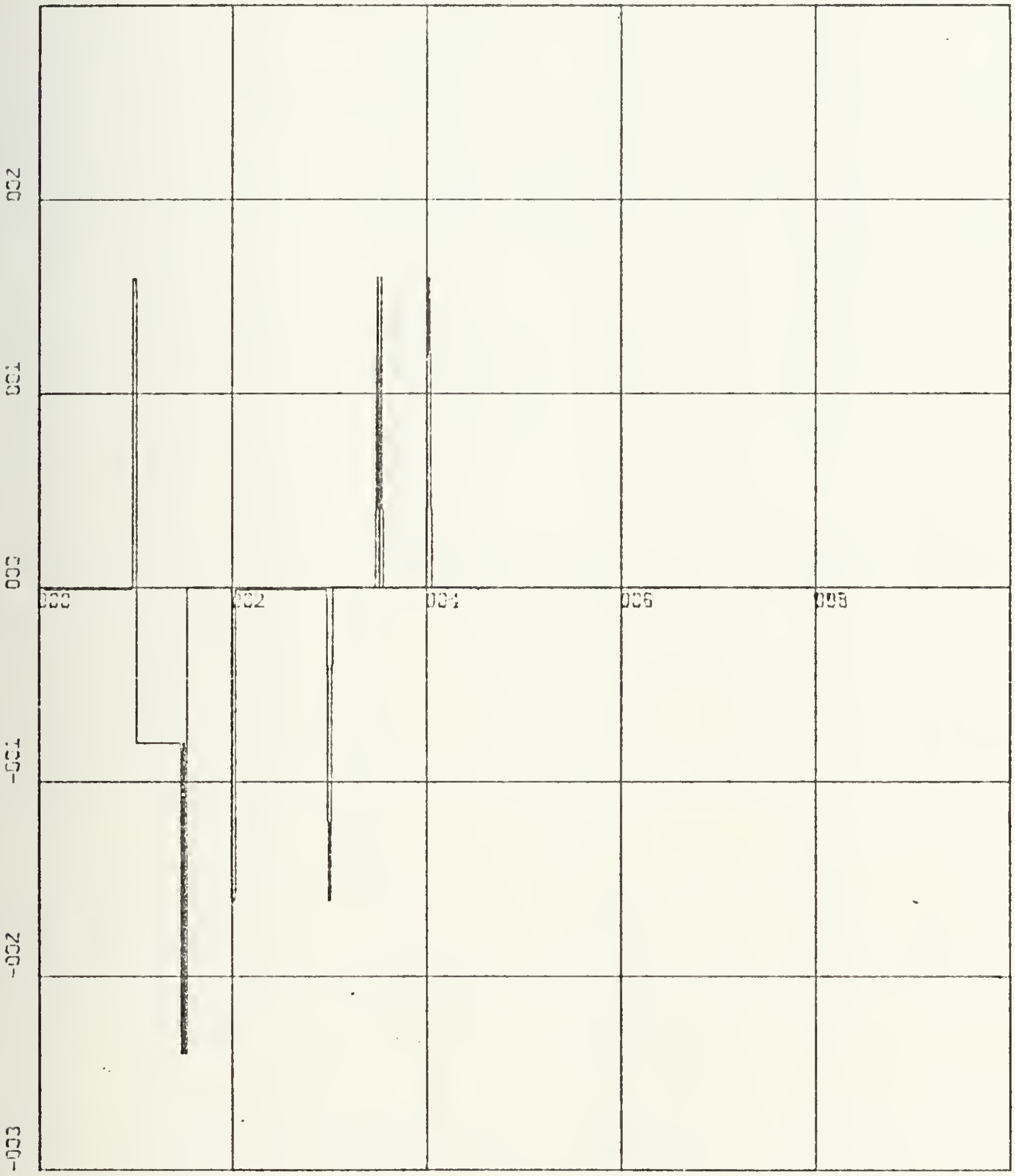
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 4, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 1 UNIT/DIV.

Figure 3-54

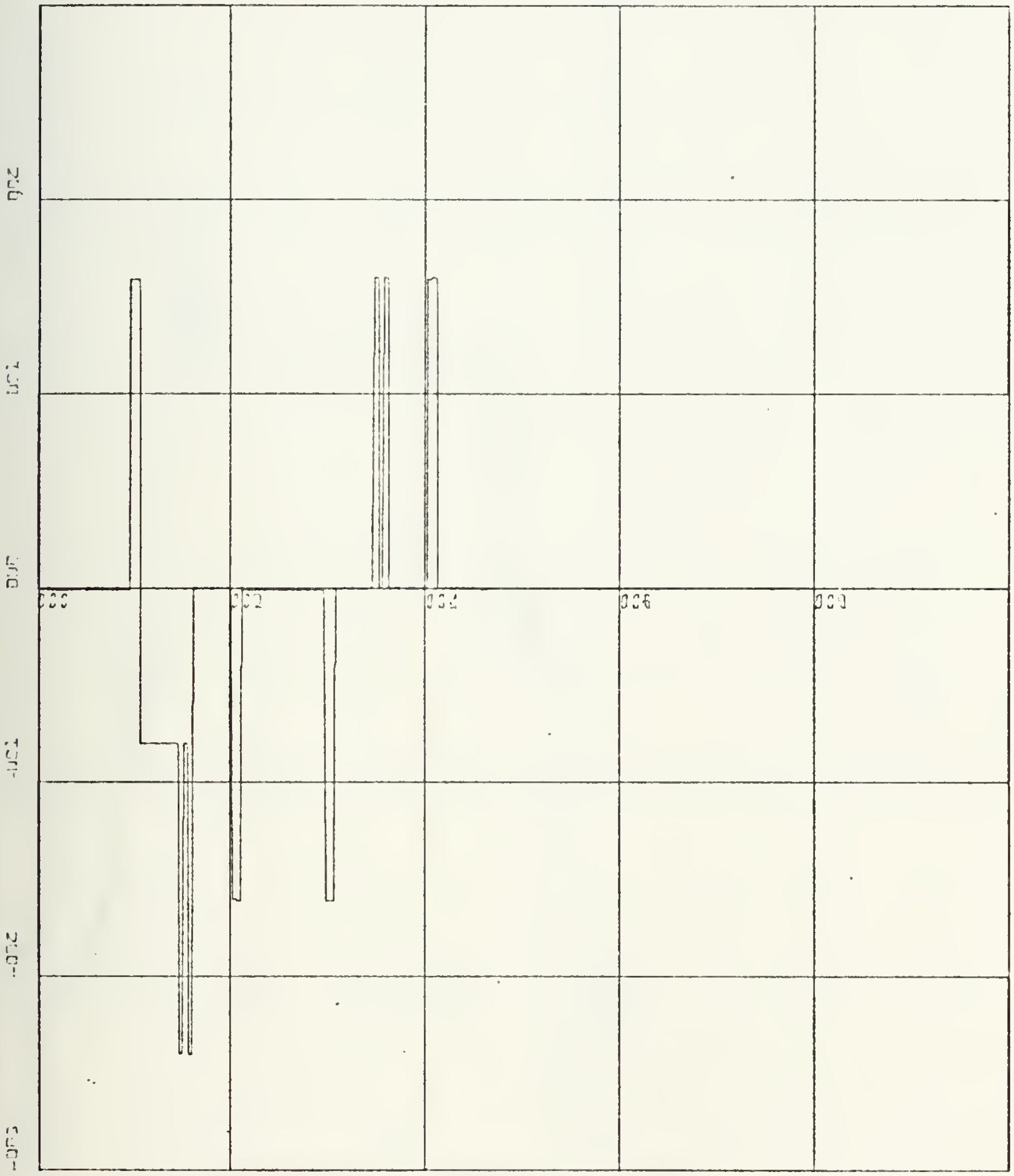
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 4, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 1 UNIT/DIV.

Figure 3-55

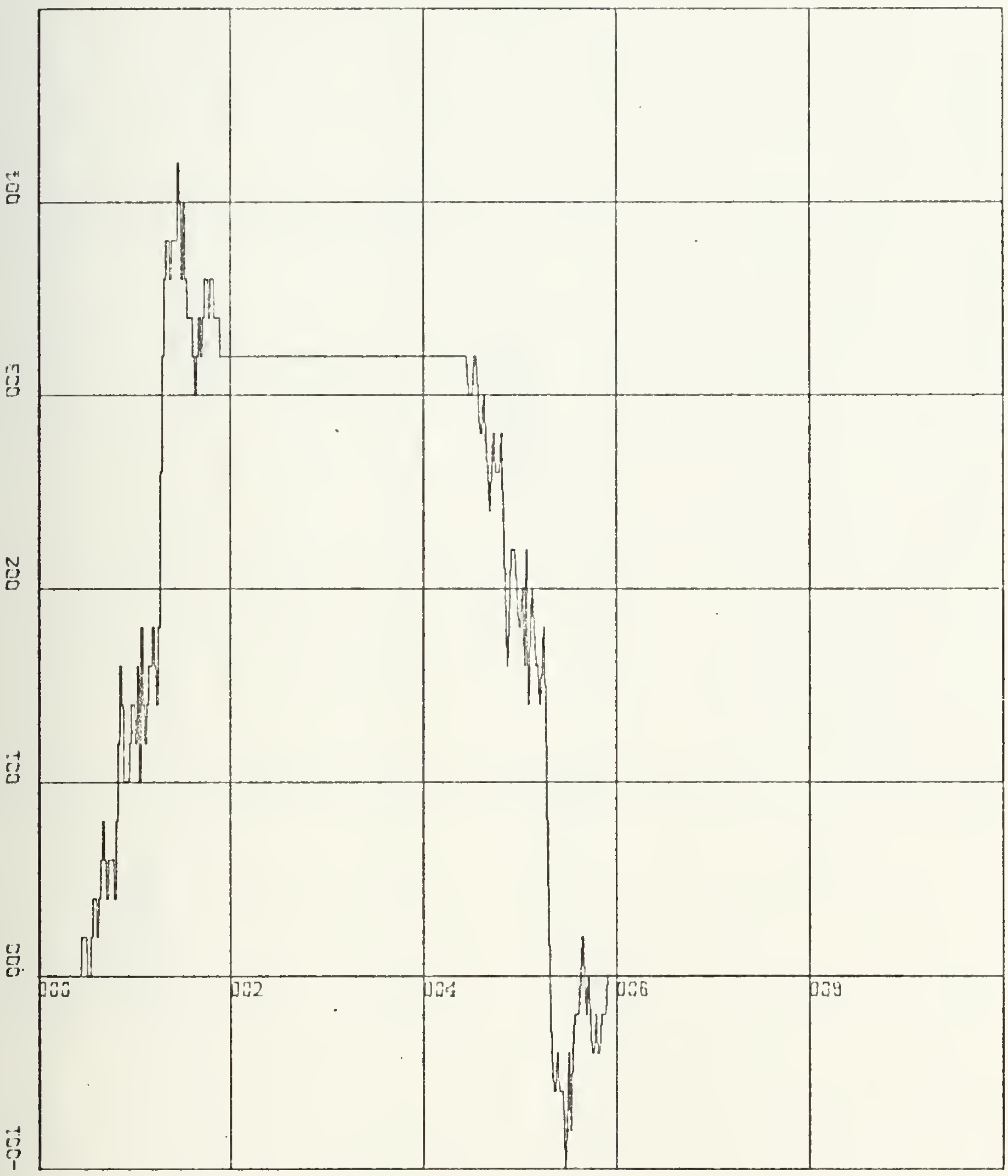
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 4, SIGNAL/NOISE = 5.0, PULSE DURATION = 2.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 200 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 1 UNIT/DIV.

Figure 3-56

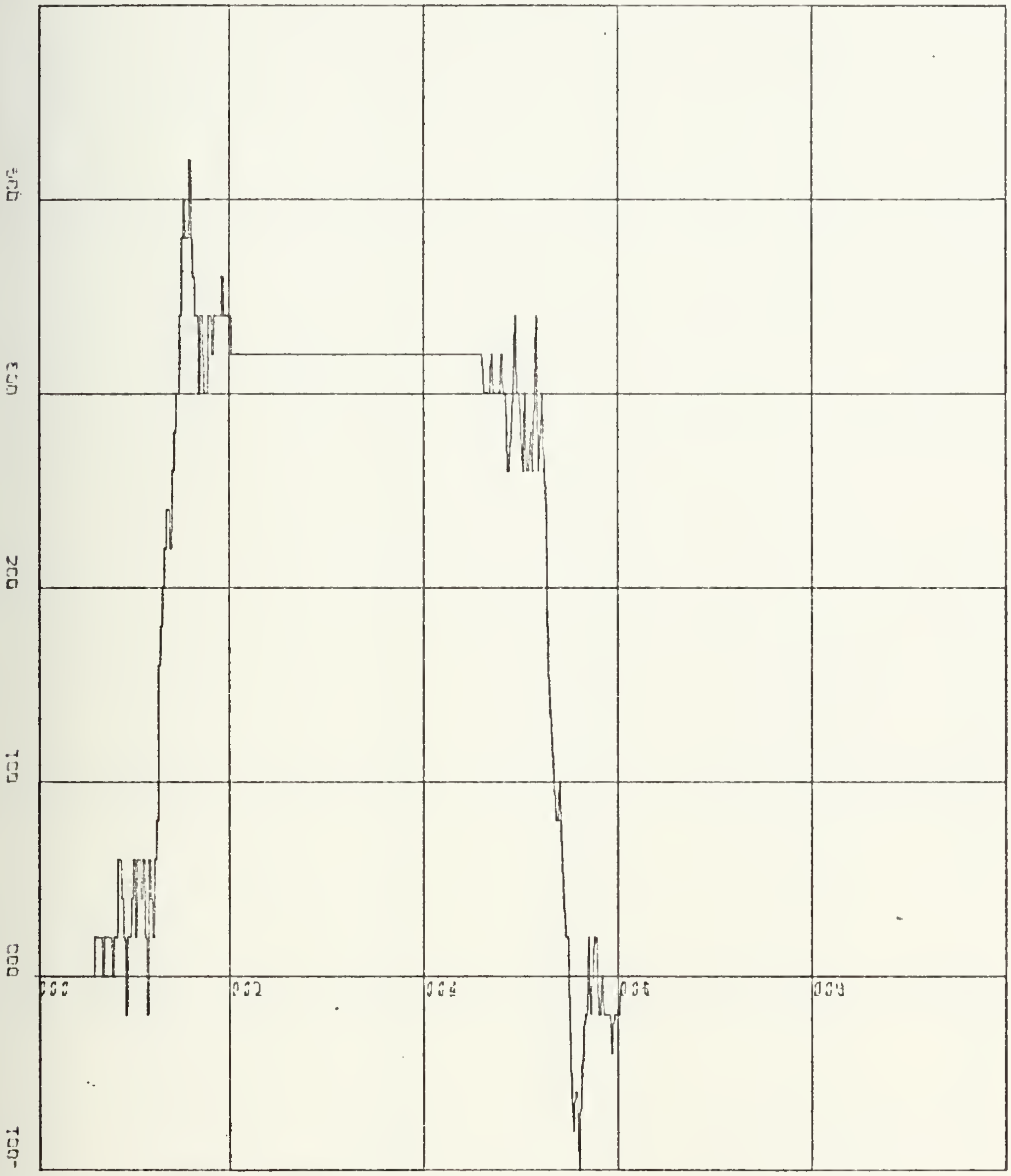
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.600 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 72.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-57

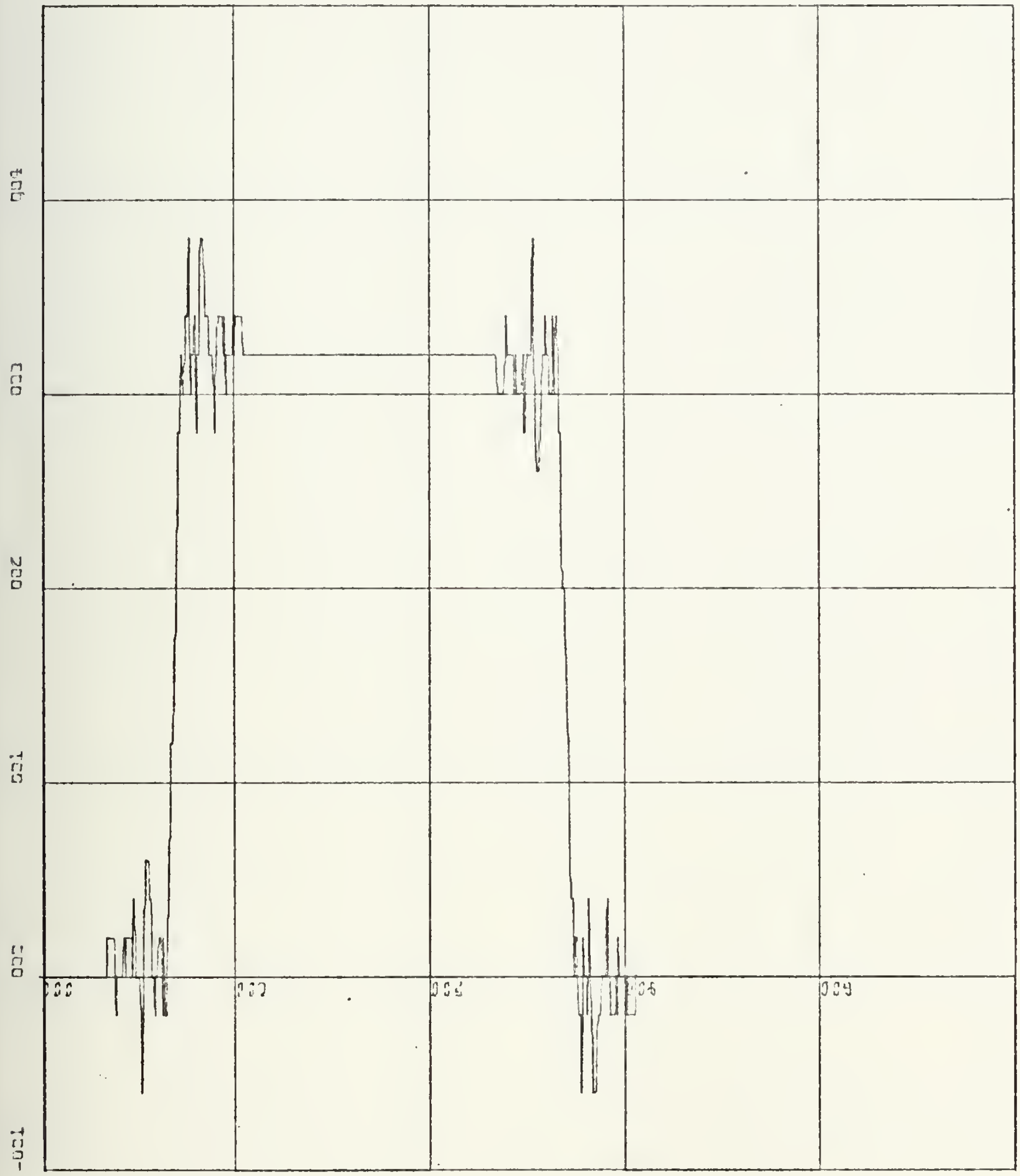
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.800 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 66.4°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-58

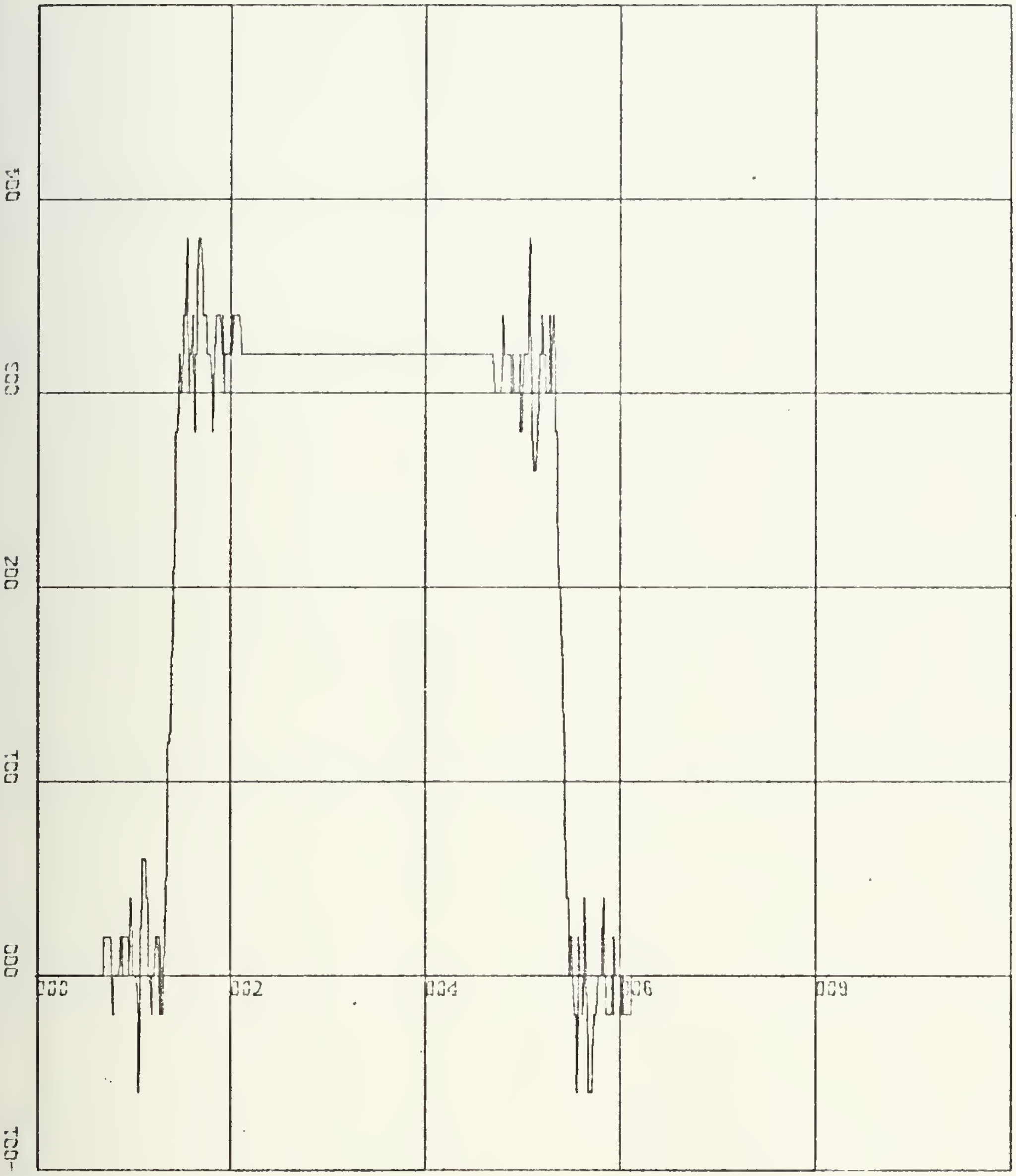
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-59

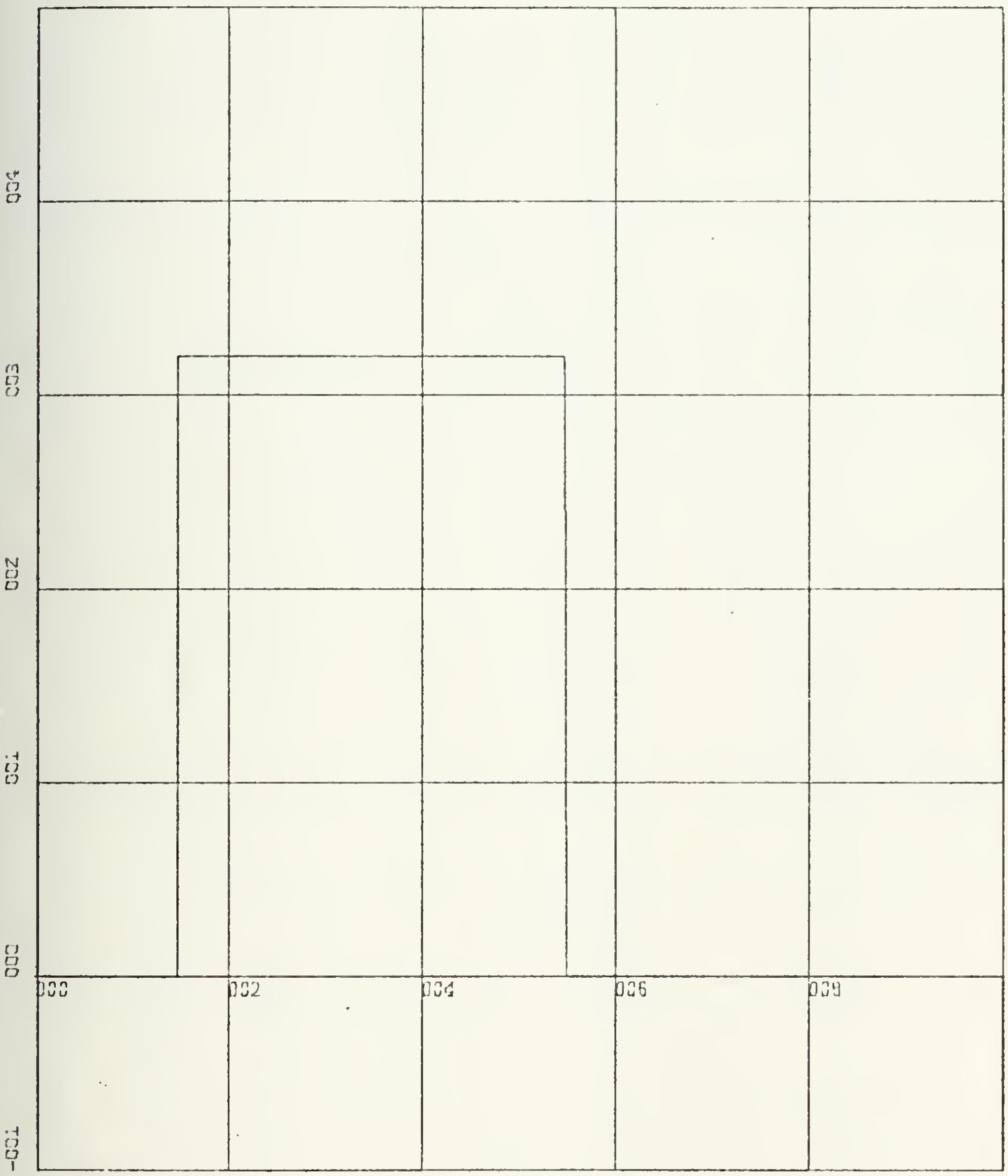
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-60

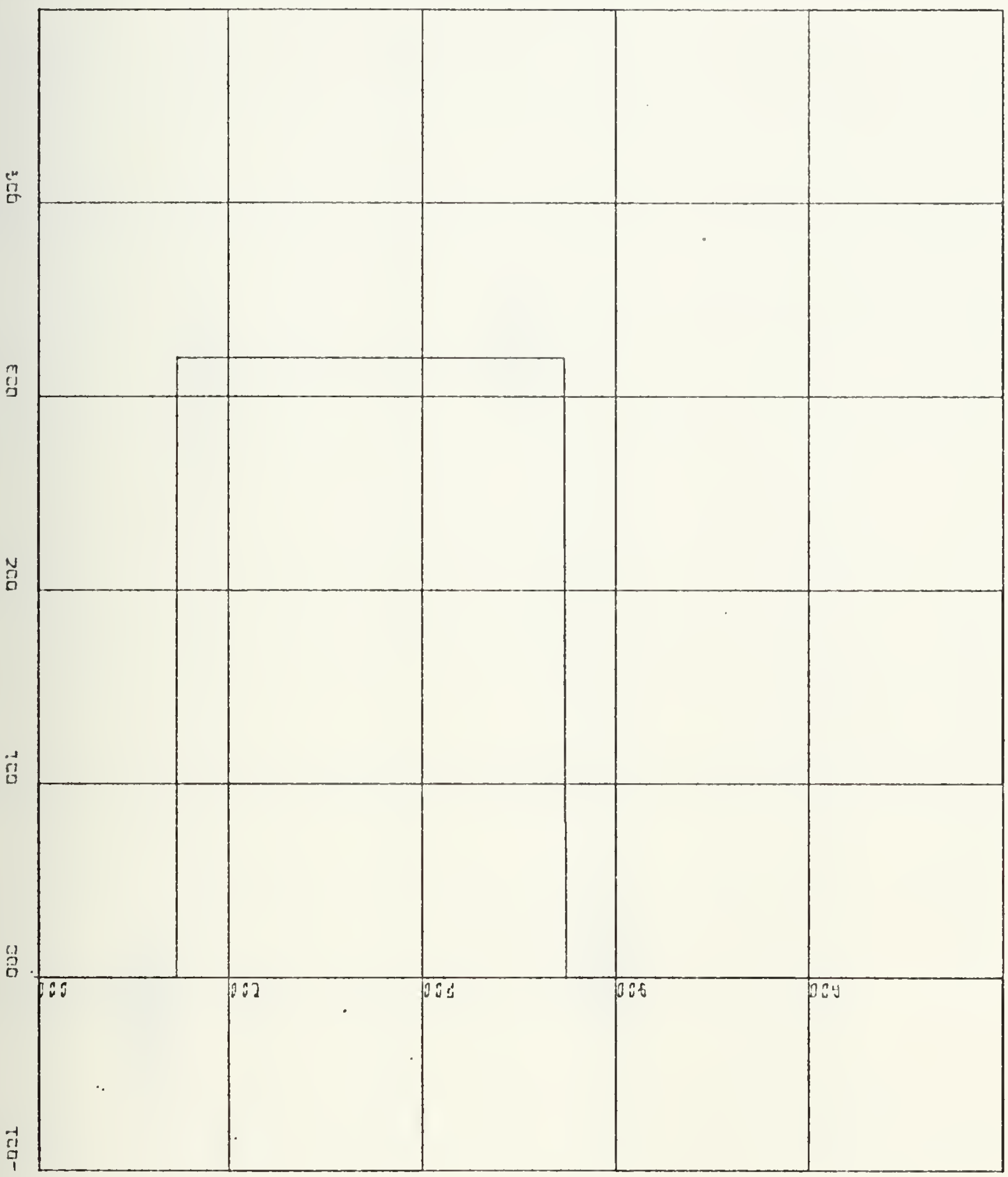
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-61

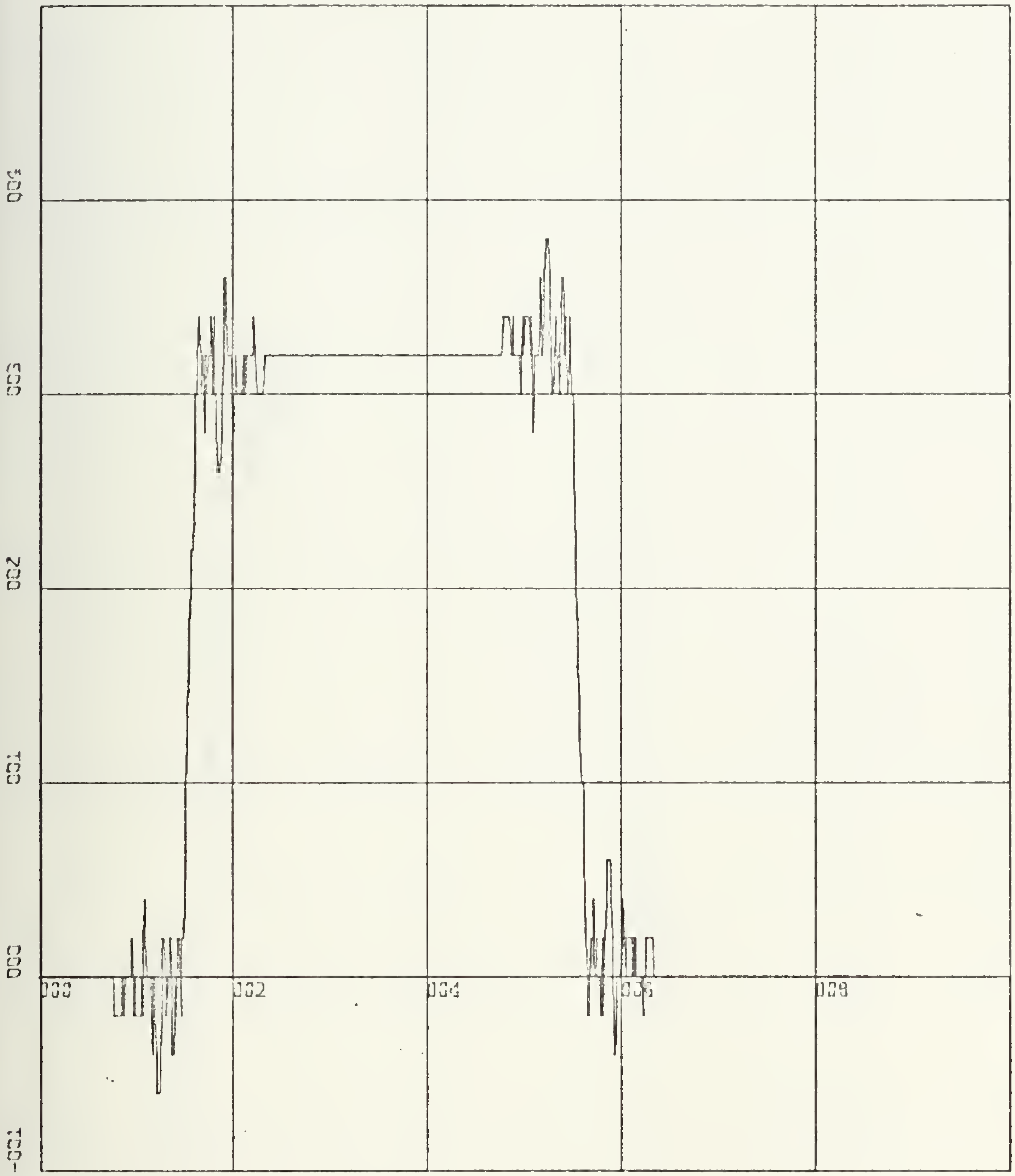
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-62

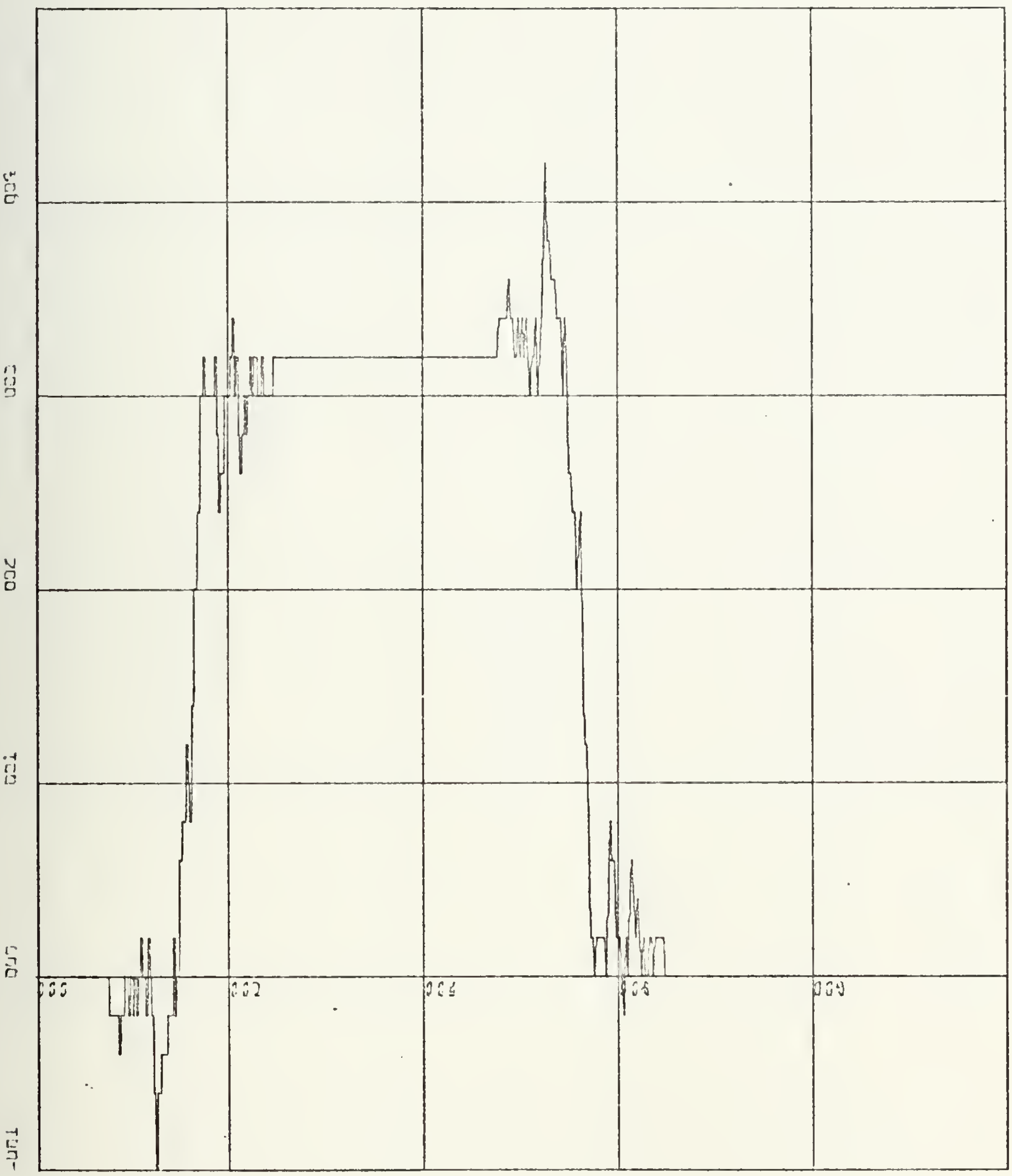
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-63

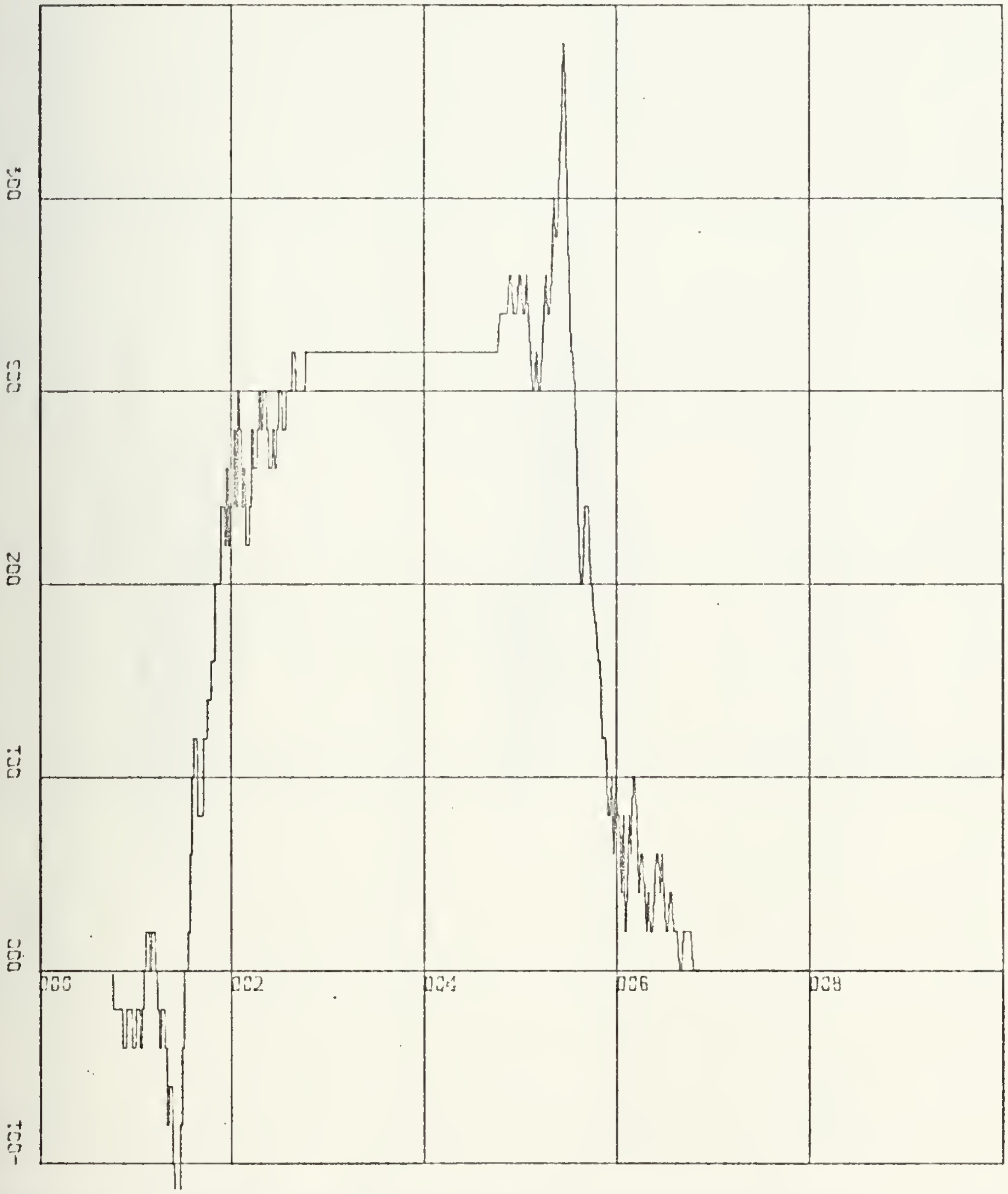
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.200 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 53.1°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-64

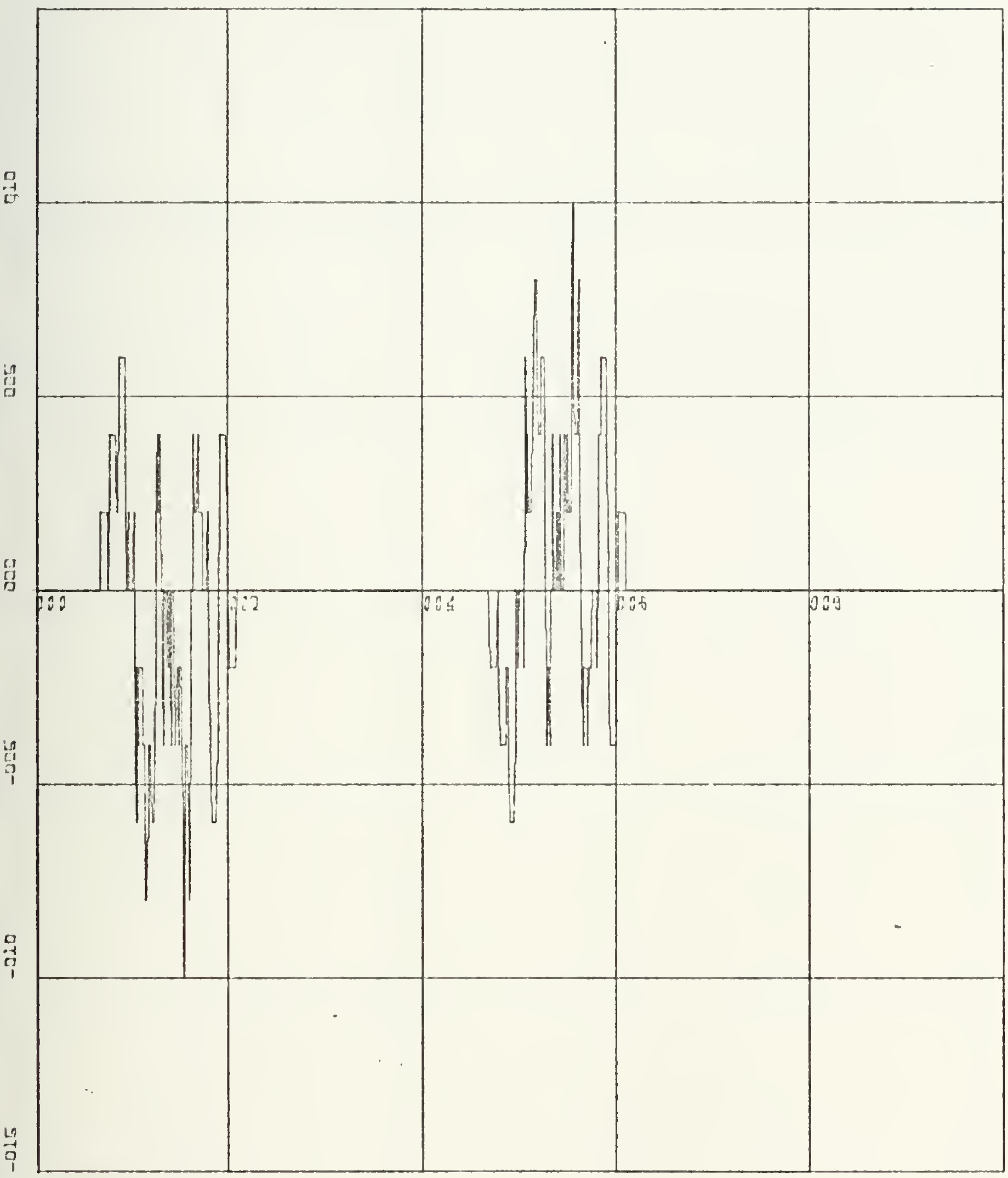
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.400 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 45.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-65

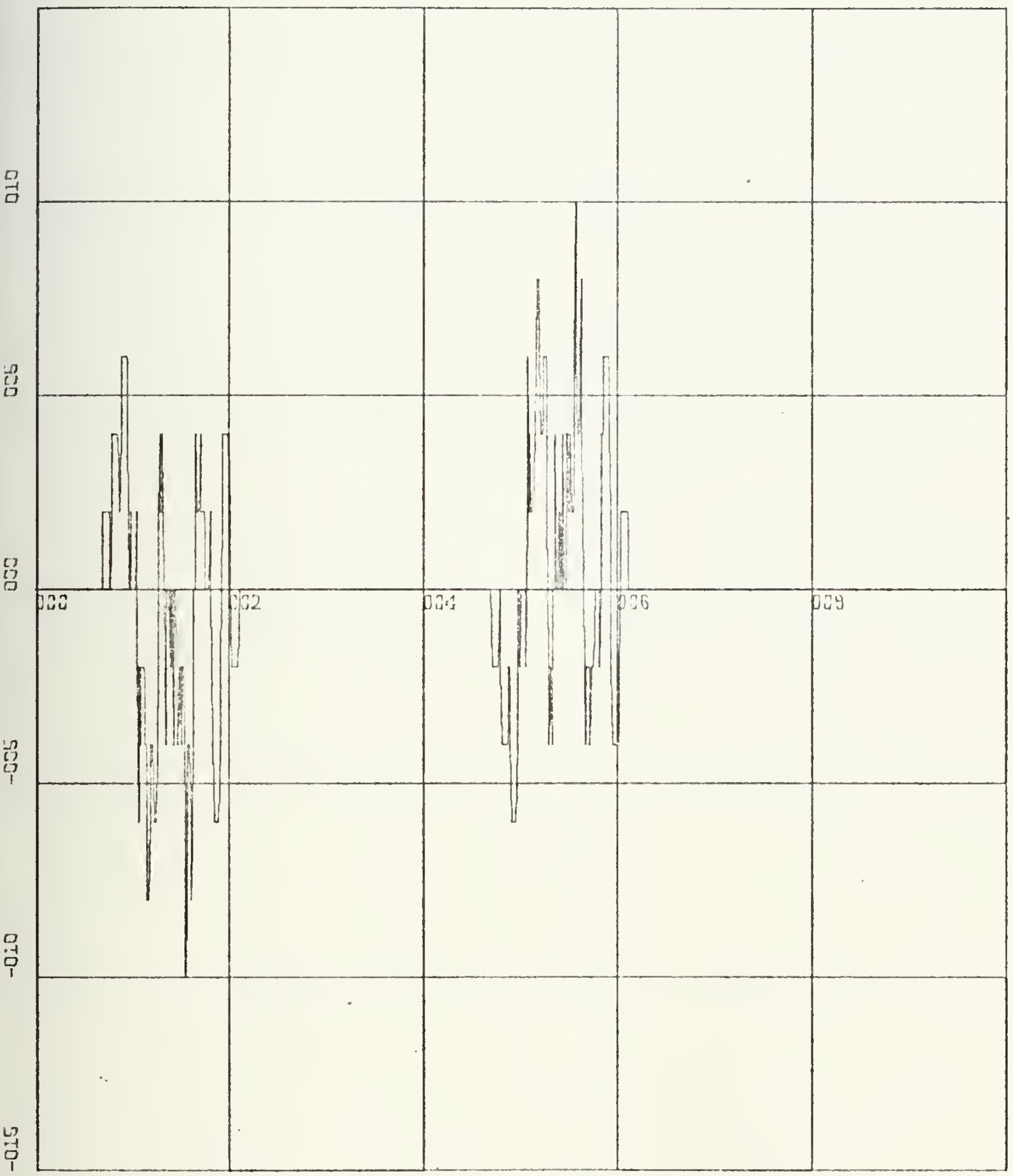
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.900 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 63.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-66

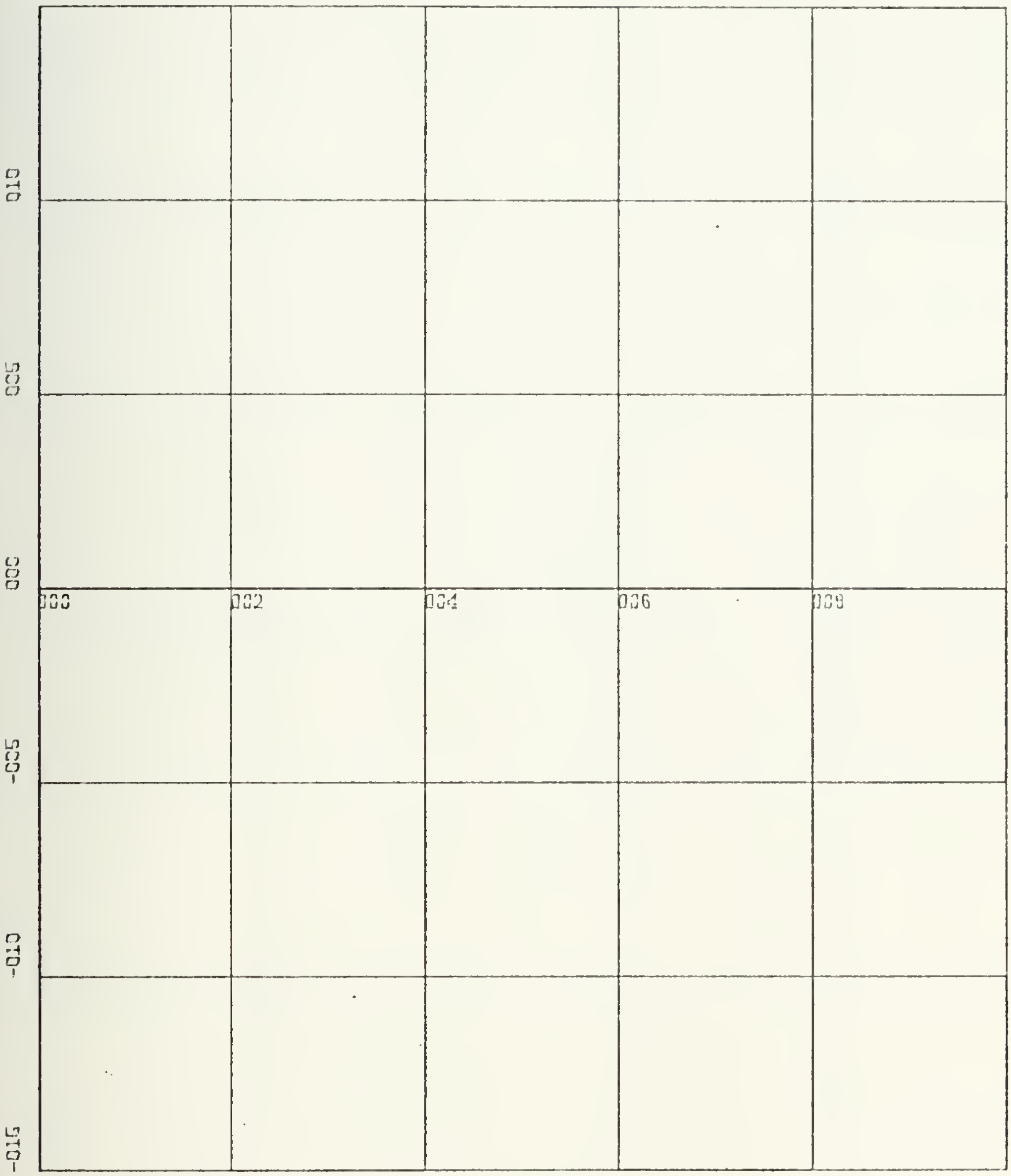
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.950 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 61.7°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-67

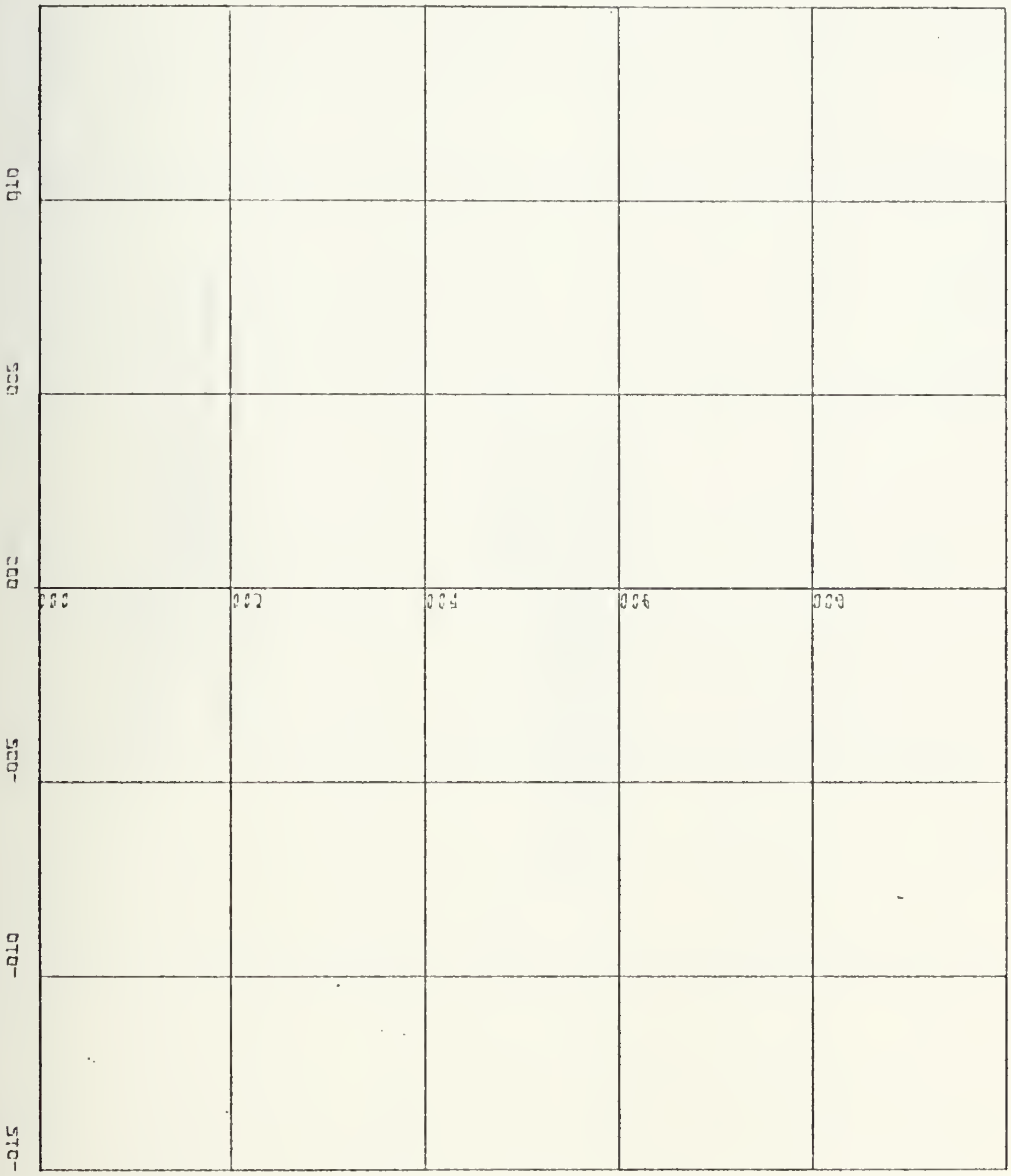
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-68

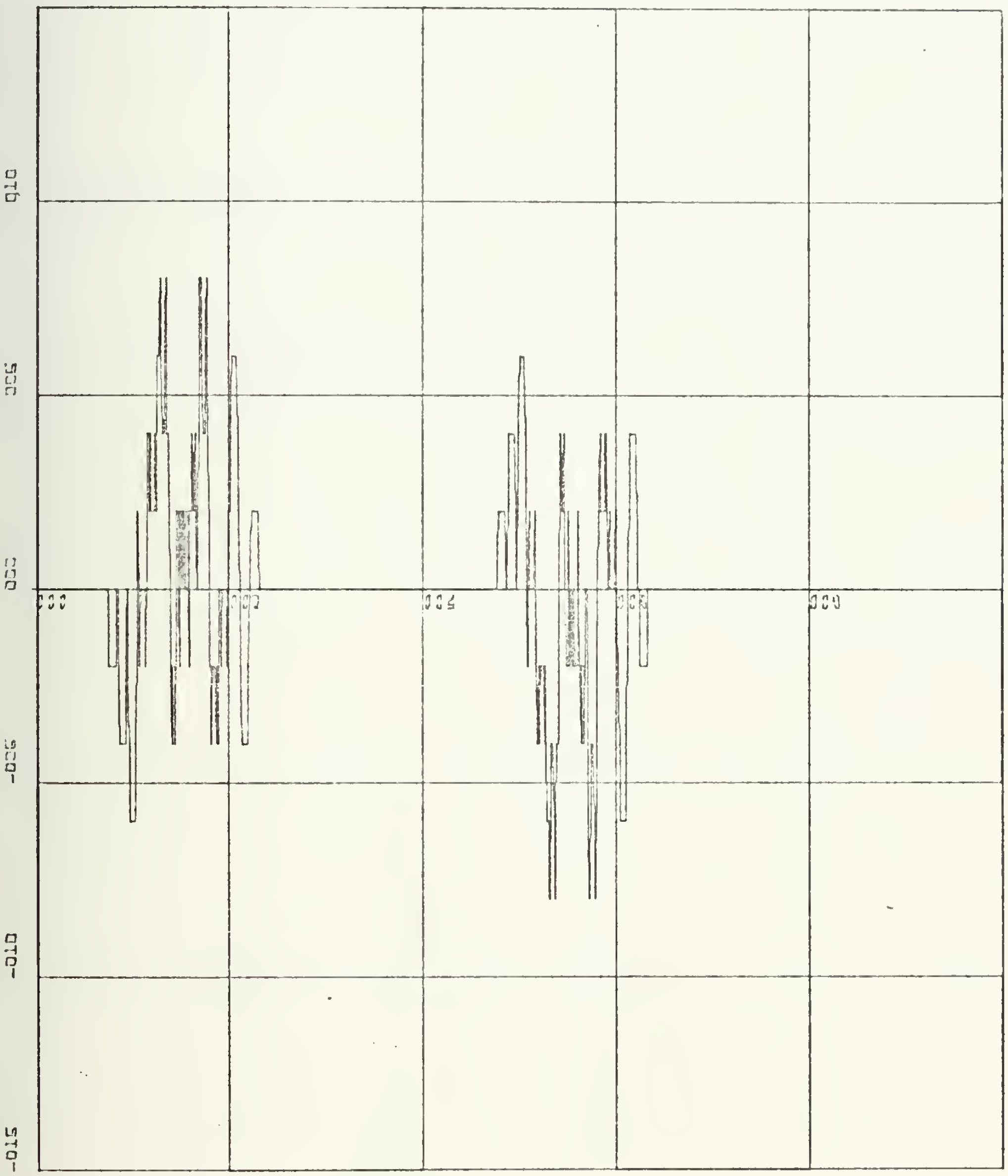
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.050 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 58.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-69

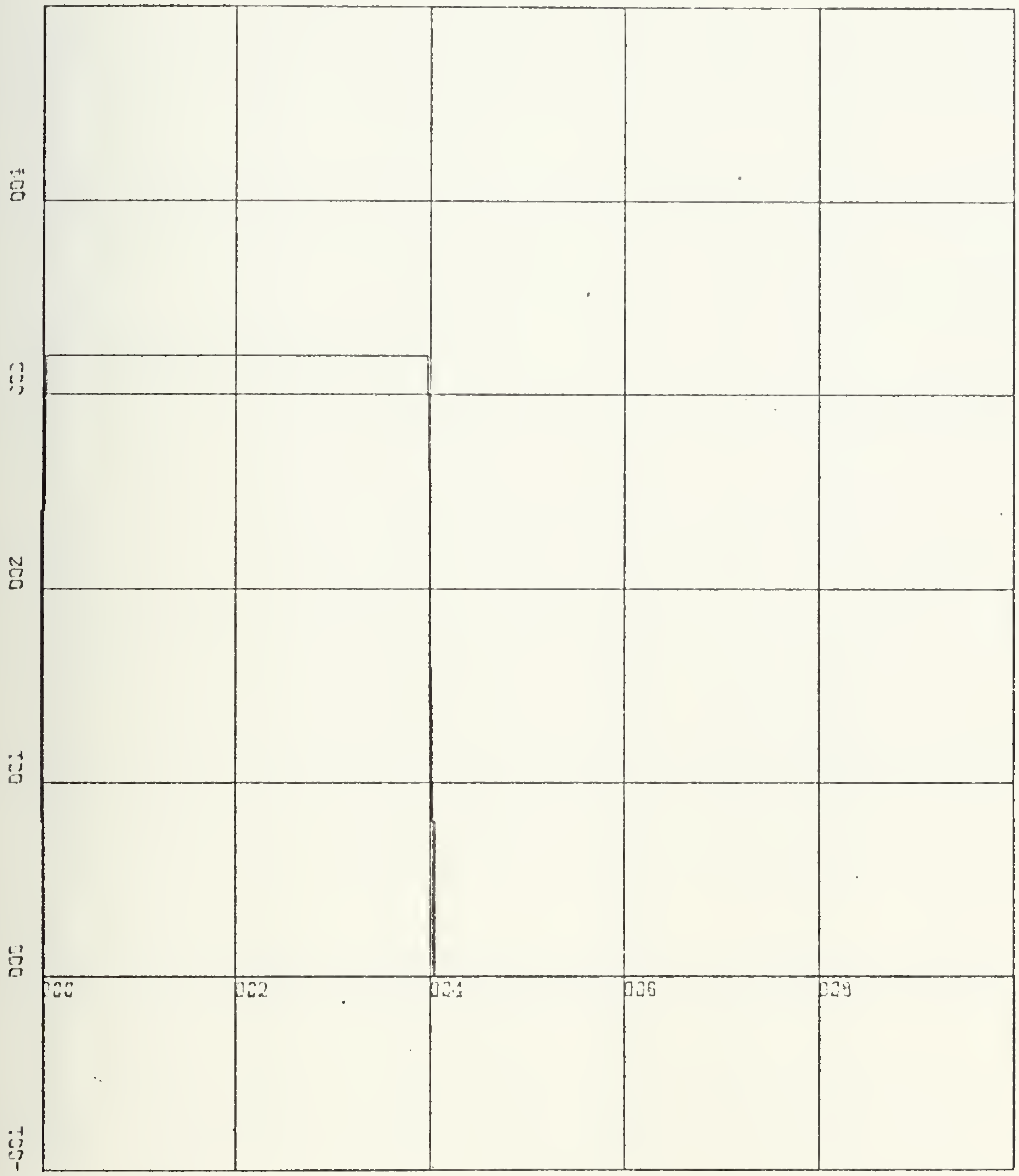
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.100 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 56.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 20.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 5 UNITS/DIV.

Figure 3-70

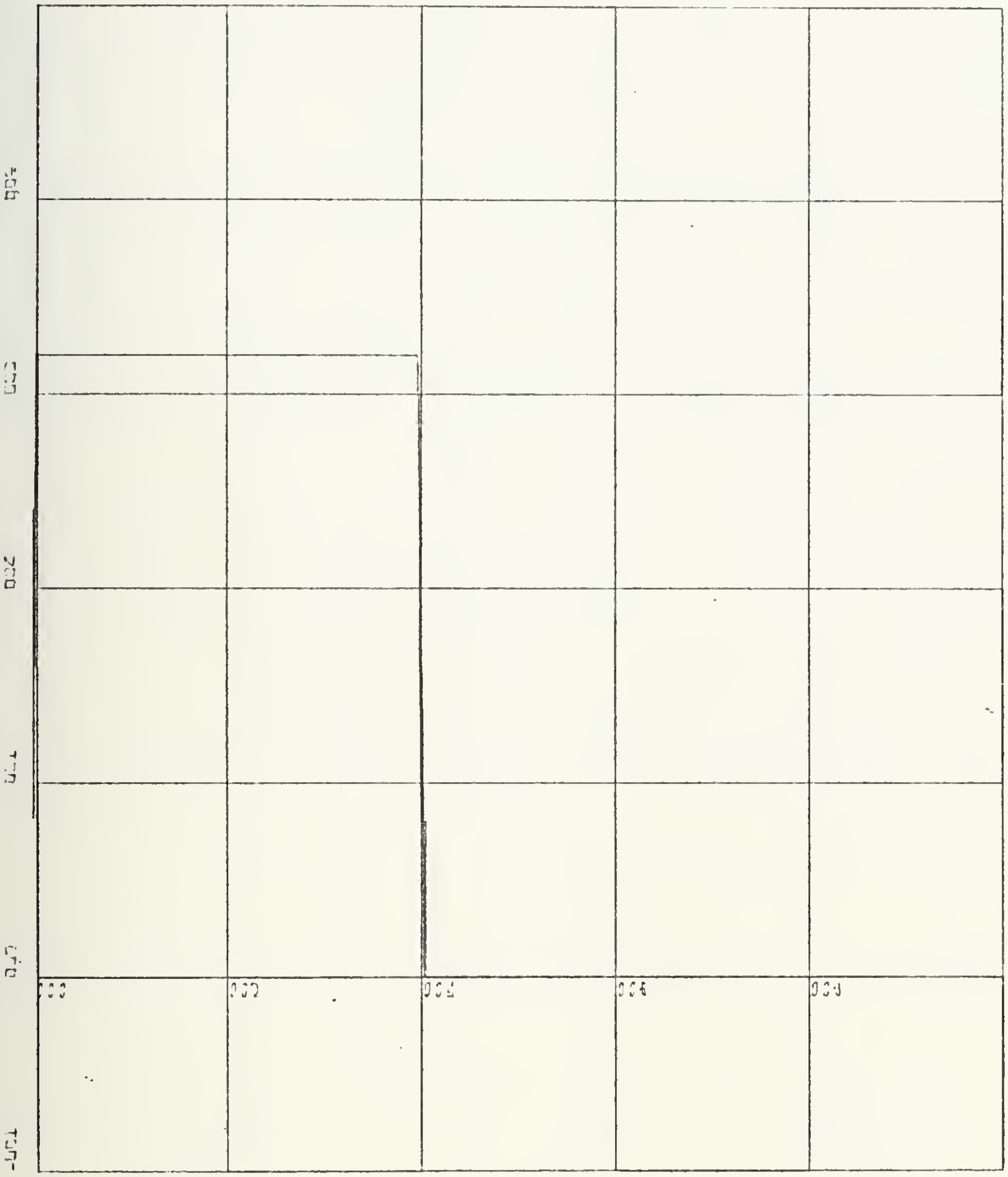
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.200 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 84.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH —— 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-71

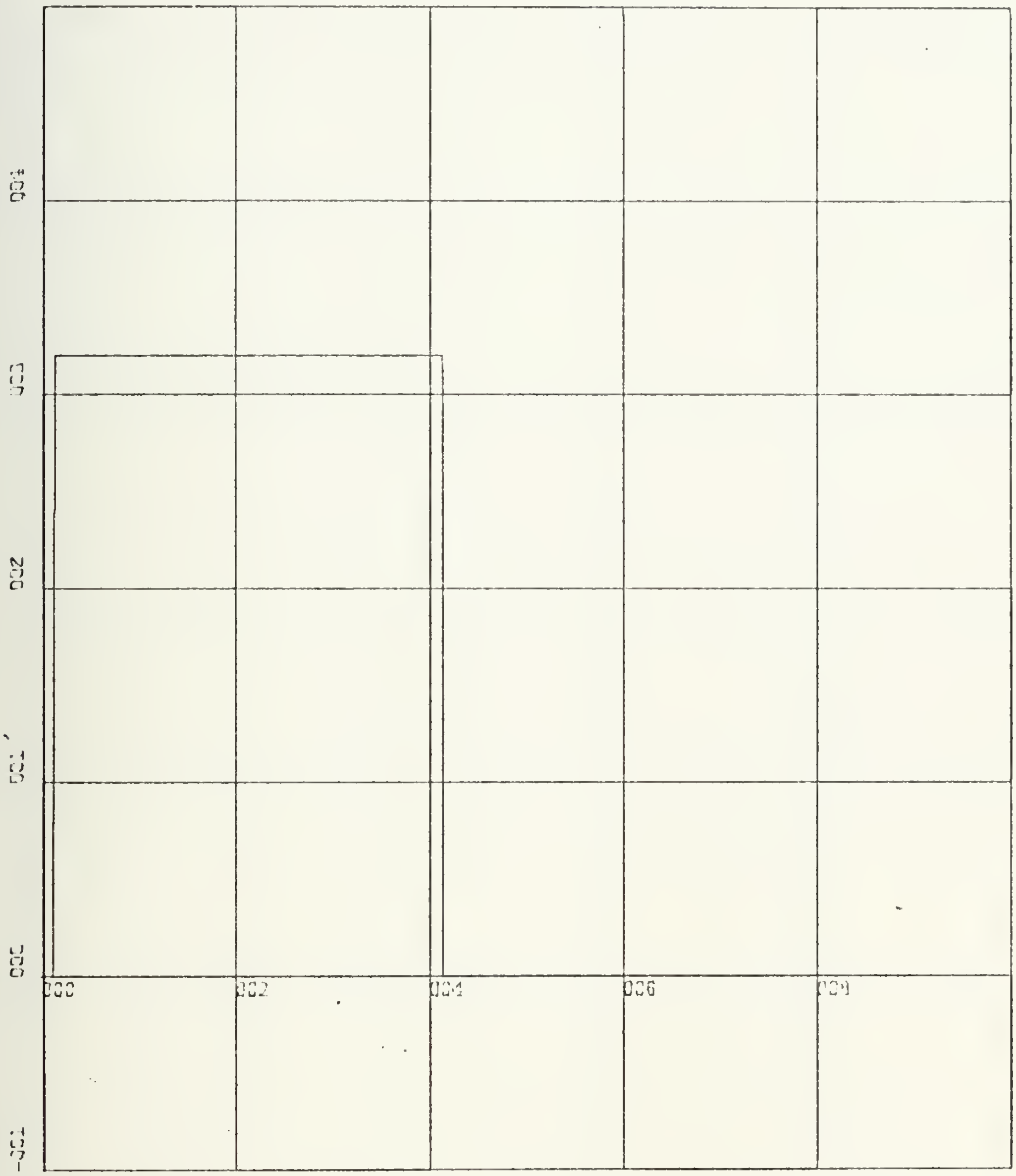
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.600 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 72.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH ---- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-72

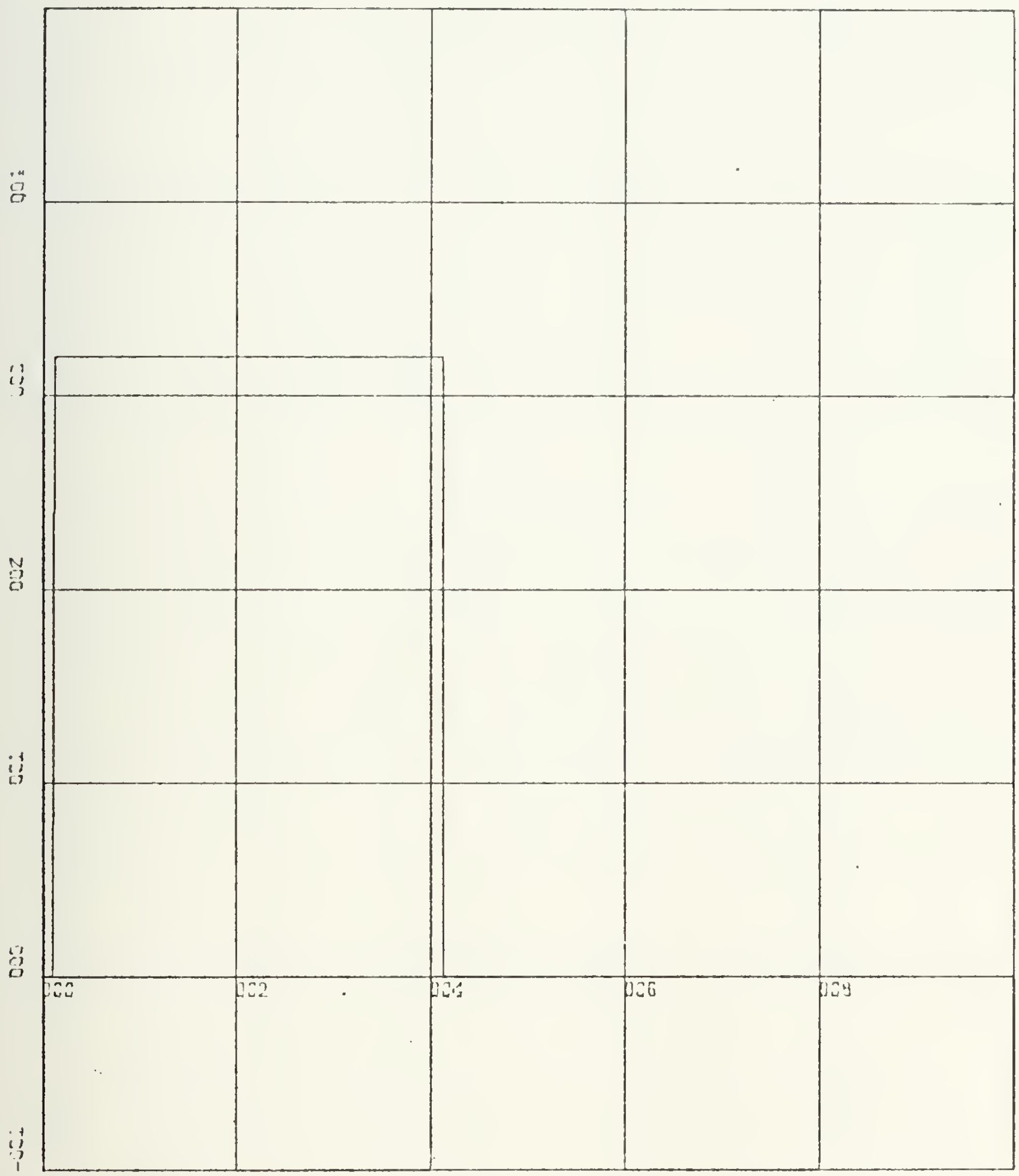
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-73

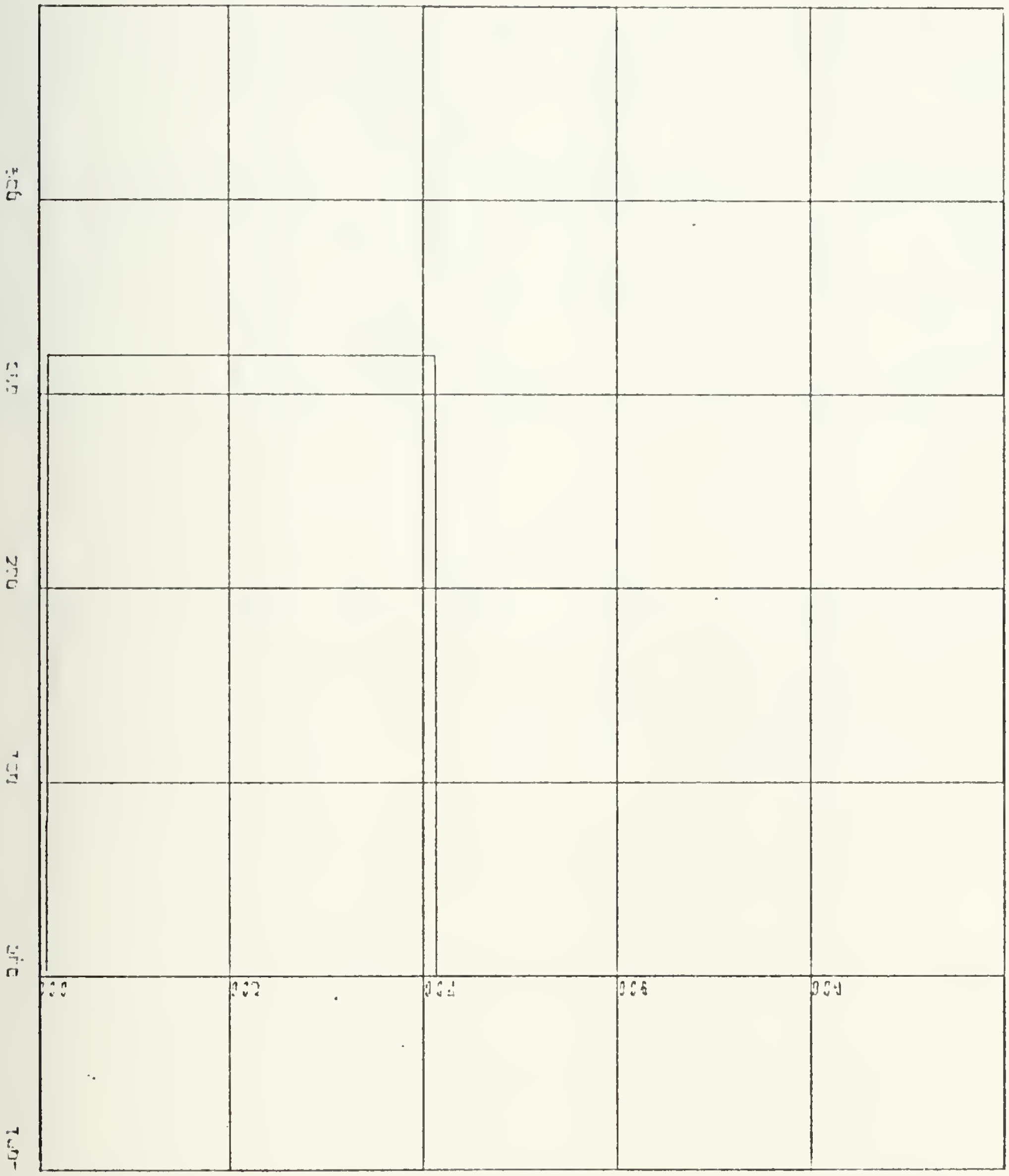
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.400 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 45.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-74

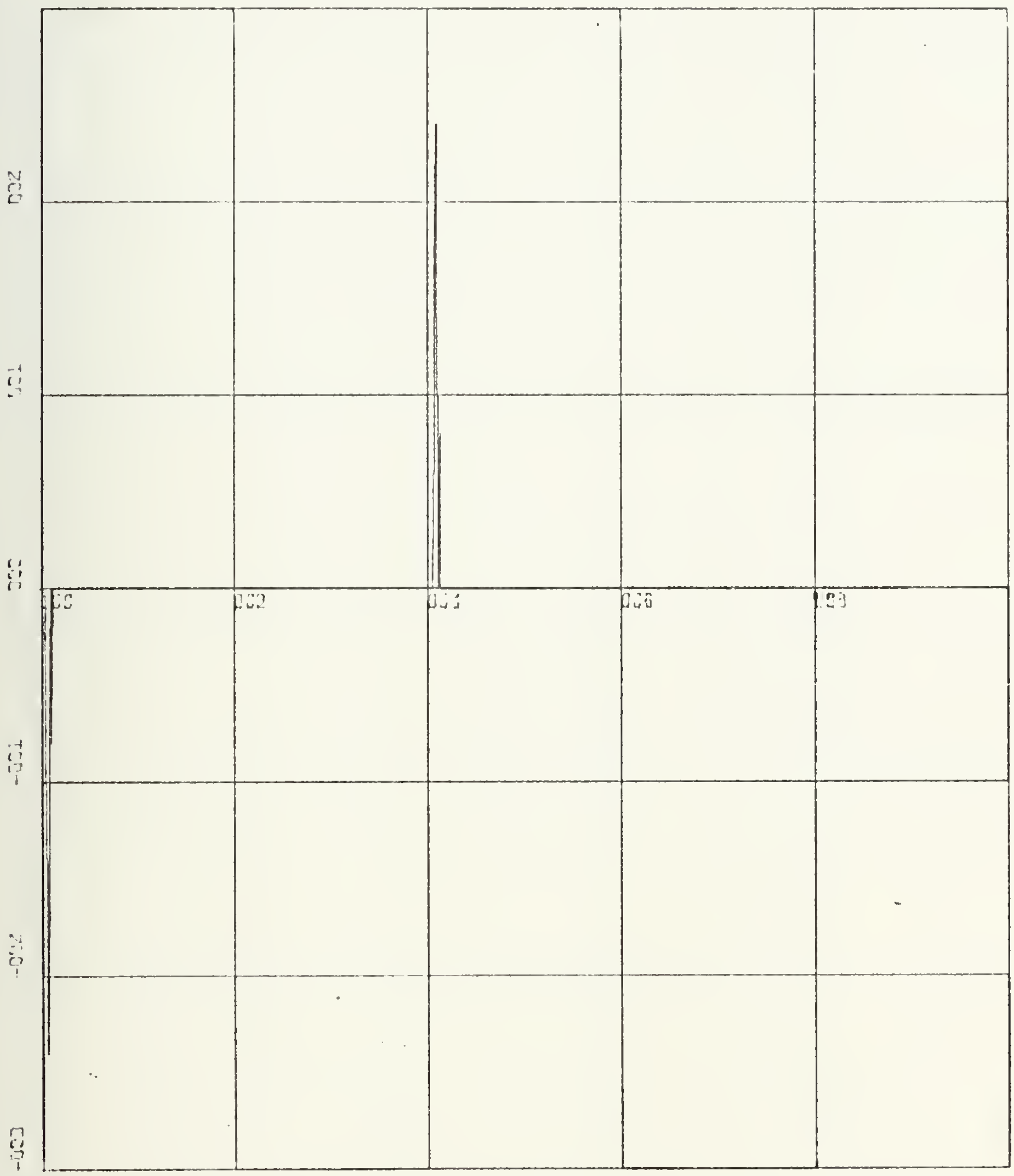
AUTOCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.800 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 25.8°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-75

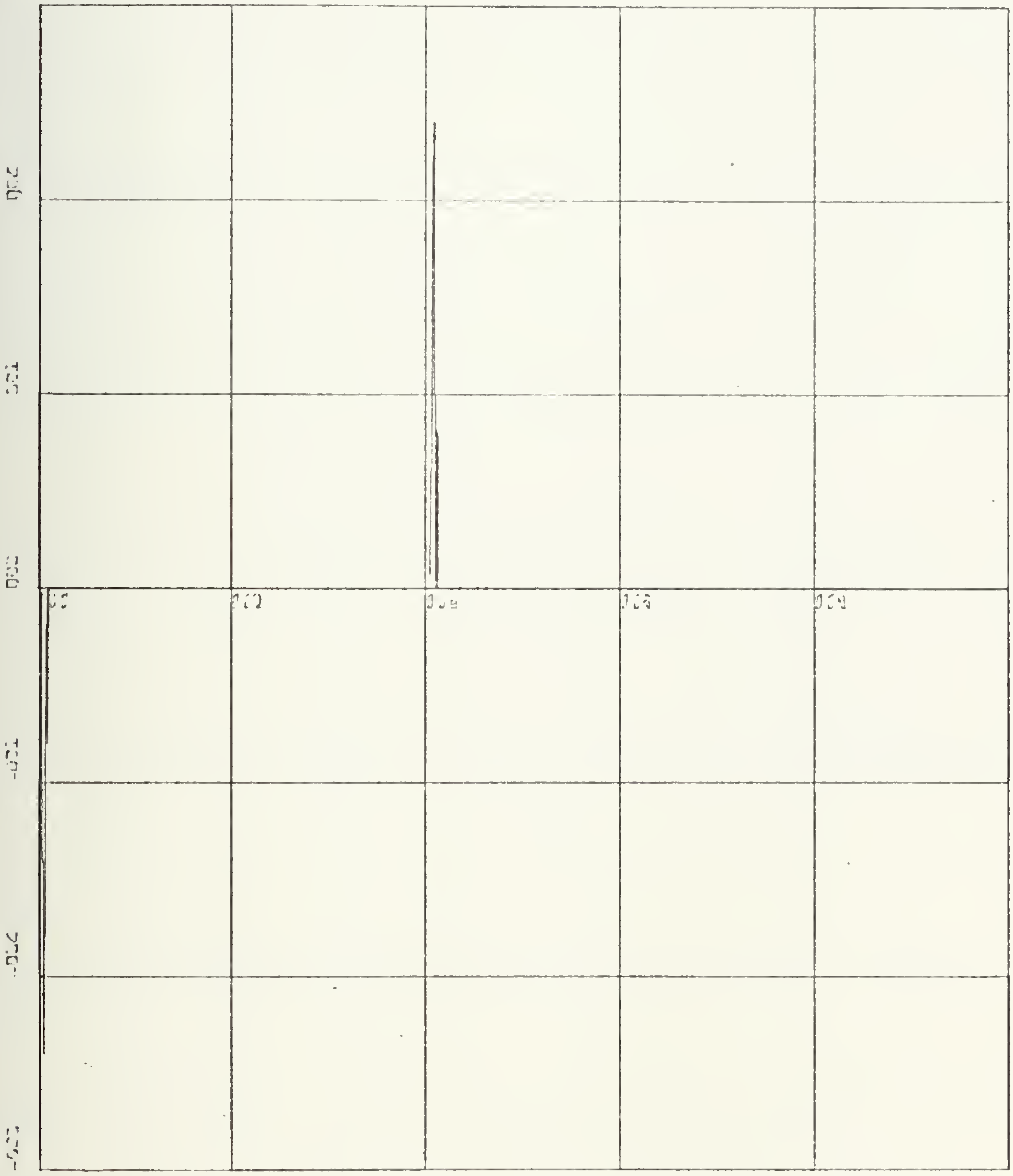
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.200 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 84.3°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-76

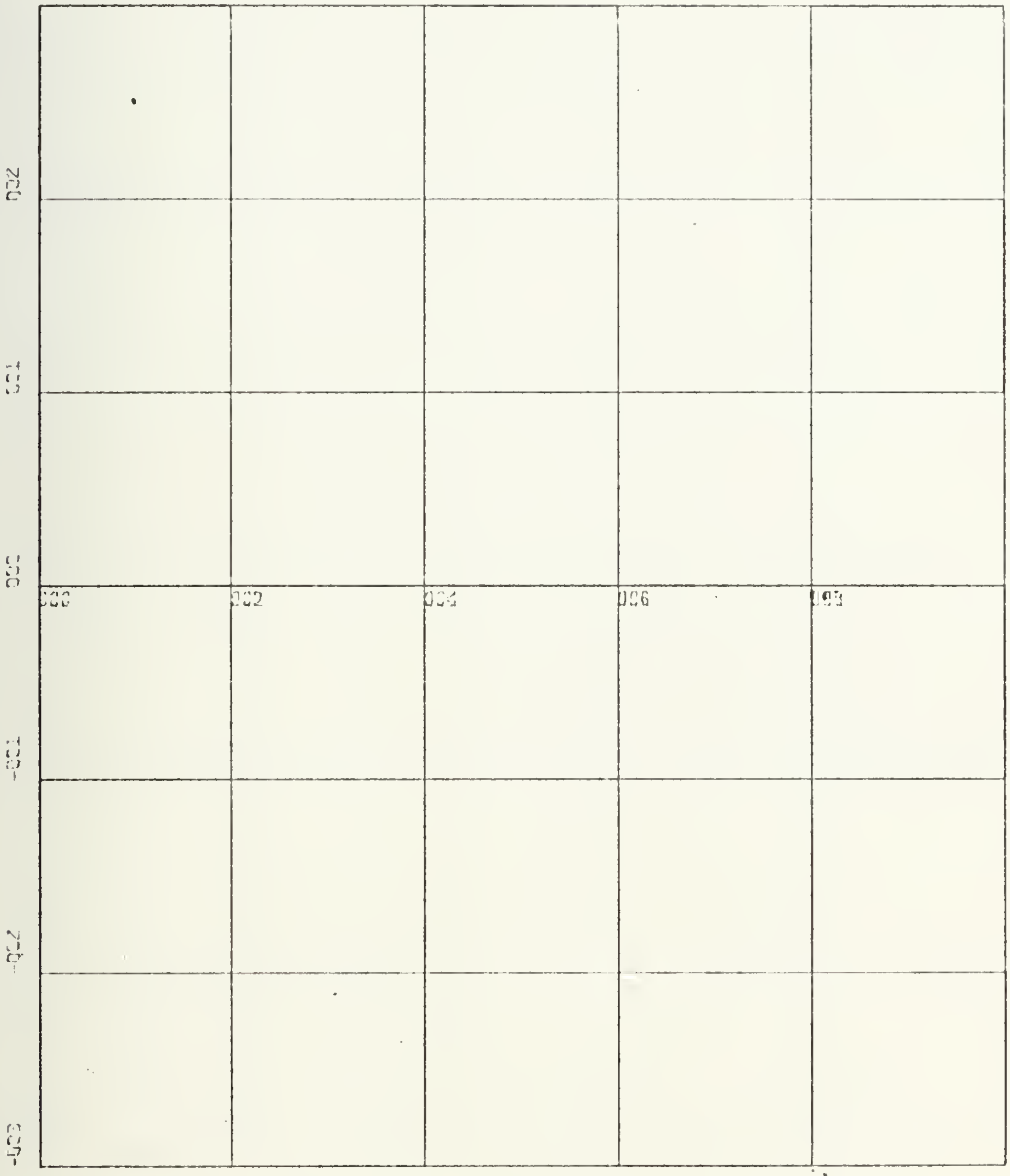
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 0.600 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 72.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-77

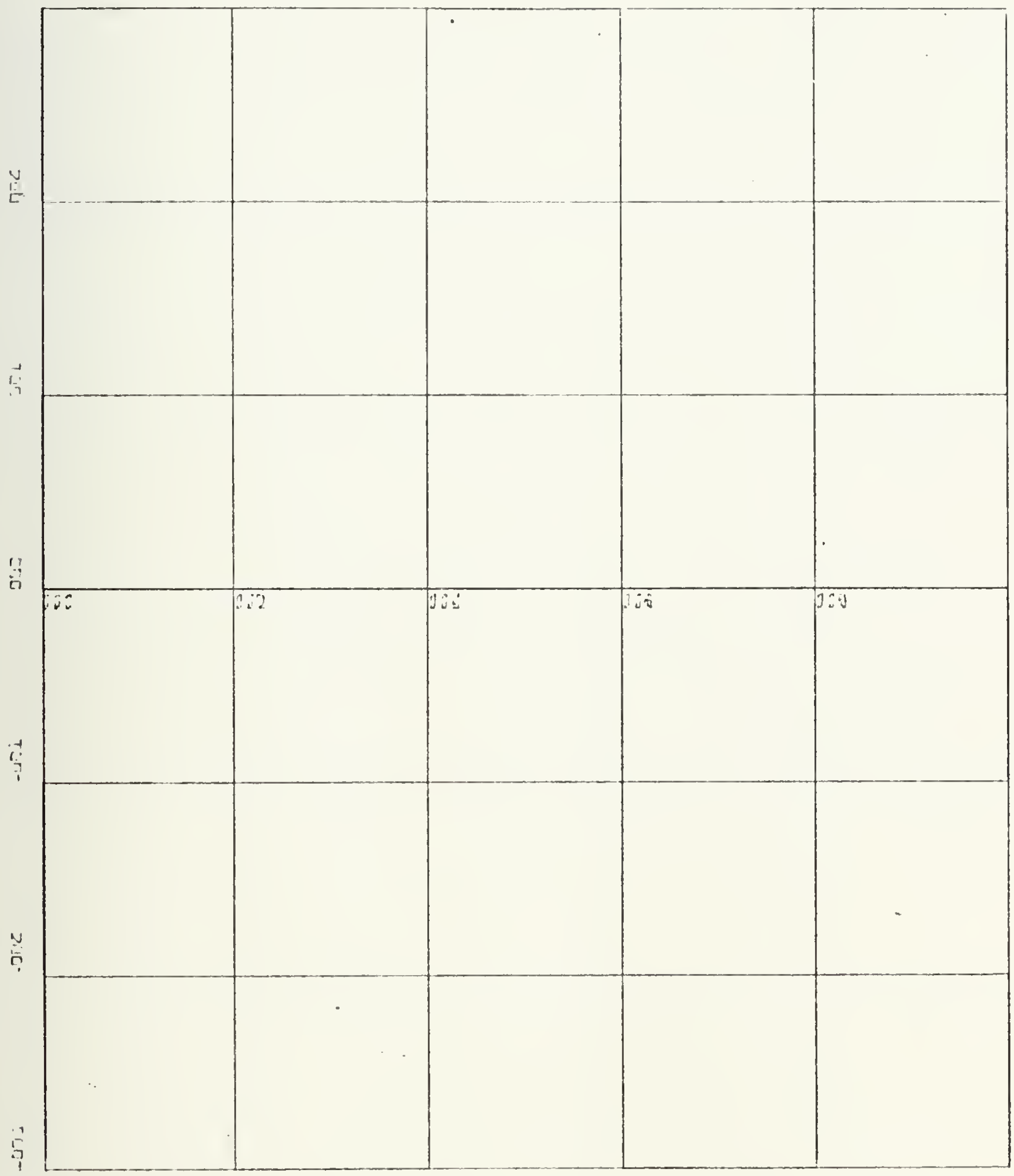
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.000 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 60.0°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-78

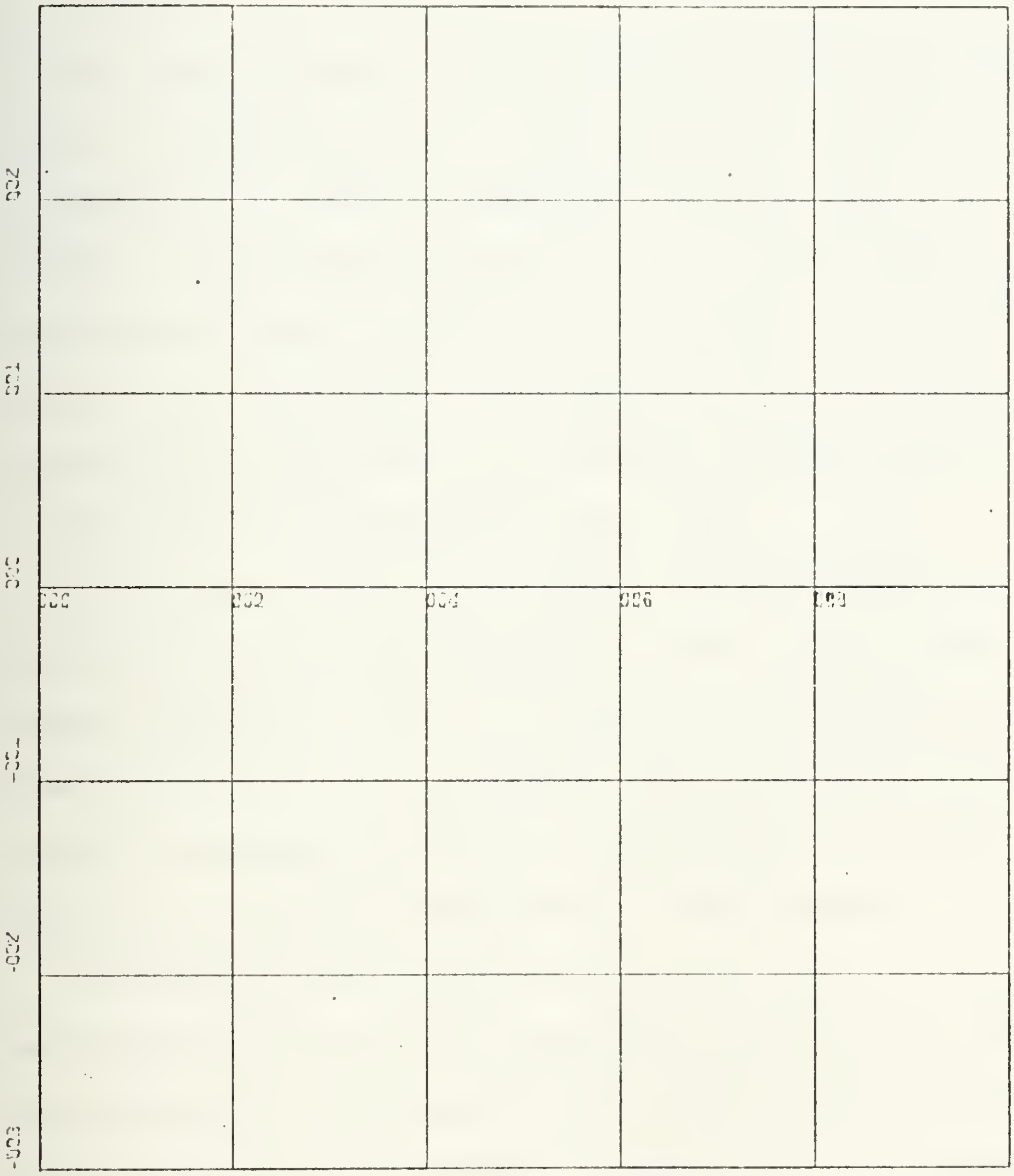
CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.400 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 45.6°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH ---- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-79

CROSSCORRELATION PLOT: AOA = 60.0°, FILTER DELAY = 1.800 TIMES THE ANTENNA DELAY (FILTER IS MATCHED FOR 25.8°), ARRAY LENGTH = 16, SIGNAL/NOISE = INF., PULSE DURATION = 200.0 TIMES THE ANTENNA DELAY.



TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE WIDTH --- 400 UNITS EQUAL ONE PULSE. AMPLITUDE UNITS ARE NORMALIZED TO PULSE HEIGHT. TIME SCALE: 200 UNITS/DIV. AMPLITUDE SCALE: 10 UNITS/DIV.

Figure 3-80

IV. SUMMARY AND RECOMMENDATIONS

A. SUMMARY

The phase coding radio direction finder using complementary sequence pairs has been evaluated with the objective of applying the basic technique to a realizable model of the system. A sinusoidal analysis of the carrier phase relationships which exist in the various portions of the signal process has shown that the unique autocorrelation and cross-correlation outputs will be produced for any arrival angle, if the system is matched for that angle.

It was shown that the inherent ambiguities of a single array system, caused by elevation angle and adjacent quadrant azimuth components, can be resolved by using a common crossed-array interferometer technique. This technique requires that a set of correlation filters corresponding to a 180° range of arrival angles be available for each array.

A computer model of the system was used to evaluate the performance of the signal processor under varying input noise conditions, and for varying array lengths and signal duration parameters. The conclusions reached from this analysis indicate that the angular resolution of the system will not be seriously affected by the array length or input signal-to-noise ratio. That is, for most input signals, the resolving capability of the system is dependent only upon the number and incremental differences between the time delay segments of the correlation filters used. However, it was also

determined that the duration of the signal model used for this analysis had a very significant effect upon the resolution. For signal pulse widths that were large compared with the time of arrival difference between antennas, the resolution of the system was seriously degraded. As a result, it was concluded that the system would not be effective against signals with this characteristic.

B. RECOMMENDATIONS FOR FURTHER INVESTIGATION

Since the analysis of the proposed system was based entirely upon the relative time of arrival of distinct modulation events, and since this was shown to place a serious limitation on the type of modulation signals that can be processed by the system, an alternate technique should be pursued. That is, the utilization of carrier phase differences, resulting from the propagation delay between antennas, might be a more acceptable parameter to use rather than time difference.

A second recommendation would be the construction of a simple scale model, utilizing one or two small arrays in order to evaluate the operation of the system in a multi-signal environment.

APPENDIX

COMPUTER PROGRAM

The computer simulation is a digital program written in FORTRAN IV level G. It was processed by an IBM 360/67 system and the built in functions are those normally available on this system.

The program requires a data deck input for the specific complementary code used in addition to the parameters required to process the code. The following words are peculiar to this program.

CODE1: First sequence in a complementary code

CODE2: Second sequence in a complementary code

SIGN: Summation sign, used for addition of correlator outputs

DELDEG: Incremental angle shift in degrees, used to obtain a new value for the angle

D: Filter delay starting value

START: The starting value in degrees of the incoming signal

STOP: The final value in degrees of the incoming signal

START-T: The starting time for the computation

STOP-T: The completion time for the computation

DSTEP: Incremental step for the parameter D, used to obtain a new value for the filter (correlator) delay

NRANTS: The number of antennas in use

SPACE: Spacing between antennas; antennas have uniform spacing

DELT: Computation time increment

TIMES: Multiplying factor used in the subprograms in order to utilize integer arithmetic and logic

JJ: Control for suppressing data printout; do not print if JJ=2

JJJ: Control for correlation function; if JJJ=1 then cross-correlate, if not then autocorrelate

JJJJ: Control to suppress the final plot which is a summary of maximum amplitude versus angle; if JJJJ=2 then do not plot

The data deck must be assembled in a specific sequence in order to have the program operate properly. Additionally if the data is not known a card must still be loaded into the data deck in order not to invalidate the read instructions. The data must be provided whether valid or not. The data cards are assembled as in Figure A-1. The first card in the data deck is CODE1 and the last is JJJJ. All data cards are punched starting in column one. CODE1 and CODE2 are formatted floating point F4.1 and each card can have up to eighteen entries. Each code is read from one card, if longer codes (greater than 18) are required then the read statement must be changed to read more than one card for each code. NRANTS, JJ, JJJ and JJJJ are formatted in integer I11 and all other data cards are formatted in floating point F11.6.

The program can provide for a pulse signal input with the pulse duration equal to any multiple of the delay due to antenna separation. The program will automatically scale the built in pulse duration (W) and the antenna separation (SPACE) to accommodate any code length. Additionally the

computation time is automatically scaled in order to include all the correlation outputs greater than zero and also provide for the maximum number of computations per angle and filter delay setting. The maximum number of computation intervals for a fixed angle and delay is 2000. The codes are read into computer memory and all necessary operations are accomplished in order to provide the proper autocorrelation and crosscorrelation functions as a function of time, angle and filter delay. If data is provided which is beyond the array storage set aside for the computation then the program will attempt adjustments in the size of the data. After twenty attempts with no success the program will terminate and print out: DATA EXCEEDS ARRAY STORAGE... JOB TERMINATED.

The program will input all parameters, compute other required parameters in accordance with the inputted parameters, compute an output for a fixed time, angle and delay and then step the time variable to the next increment. After the time variable has been stepped across its range the filter matching coefficient (D) is incremented to a new value, and the output is computed for all time values as the time variable is again swept through its range. After the matching coefficient has been swept through values from its initial value to a final value equally spaced above D=1.000, the angle variable may then be incremented to a new value. The outputs are then computed across the time range and the angle range. In this manner the outputs of the filters

(correlator outputs) are computed for all times of interest, all angles of interest and all desired variations of the filter delay lines.

The outputs available from this program are printer plots and fine grain continuous plots of correlation output versus time for each and every value of angle and filter delay. The printer plots contain all pertinent parameters and scaling information.

A listing of important parameters and statements concerning the interpretation of the various control commands are printed at the beginning of each computer run to aid in analysis of the data and the plots. The maximum core size is 150K and, due to the fine grain computation required by the program, execution time for most investigations is in the 15 to 20 minute region.

TYPICAL DATA DECK

```
1.0 1.0 1.0-1.0 1.0 1.0-1.0 1.0 1.0 1.0 1.0-1.0-1.0-1.0 1.0-1.0
1.0 1.0 1.0-1.0 1.0 1.0-1.0 1.0-1.0-1.0-1.0 1.0 1.0 1.0-1.0 1.0
  1.000000
  3.000000
  0.990000
 60.000000
 60.000000
  0.000000
  0.000000
        16
         2
         2
         2
```

```
CARD 1...FIRST SEQUENCE OF THE CODE PAIR.....(CODE1)
CARD 2...SECOND SEQUENCE OF THE CODE PAIR....(CODE2)
CARD 3...SUMMATION SIGN......................(SIGN)
CARD 4...ANGLE STEP..........................(DELDEG)
CARD 5...FILTER DELAY STARTING VALUE.........(D)
CARD 6...ANGLE START.........................(START)
CARD 7...ANGLE STOP..........................(STOP)
CARD 8...COMPUTATION TIME START..............(START-T)
CARD 9...FILTER DELAY STEP...................(DSTEP)
CARD 10..NUMBER OF ANTENNAS..................(NRANTS)
CARD 11..CONTROL FOR DATA PRINT..............(JJ)
CARD 12..CONTROL FOR TYPE OF CORRELATION.....(JJJ)
CARD 13..CONTROL FOR OUTPUT VS ANGLE PLOT....(JJJJ)
```

FIGURE A-1

SIMULATION OF RADIO DIRECTION FINDER UTILIZING GOLAY PHASE-CODED SIGNAL PROCESS

```
      COMMON/ZZZ/ADD(2000)
      COMMON/AAA/ADD1(2000)
      COMMON/BBB/ADD2(2000)
      COMMON/RRR/TA(200),TOUT(200),SUM(50),CODE1(18),CODE2(18),
     1PT(2000),POUT(2000),KDUMB(18),JDUMB(18),TIME(400),
     2IDUMB(18),SIG(50)
      REAL*4 IDUMB,JDUMB,KDUMB
      REAL LABEL/'    '/
      REAL*8 TITLE(12)/'PRISAZNK',11*'        '/
      INTEGER*4 T3,FIL3,DELT3,W3,D3
      DO 7 I=1,18
      CODE1(I)=0.0
      CODE2(I)=0.0
      JDUMB(I)=0.0
      KDUMB(I)=0.0
    7 CONTINUE
C
C
C     READ IN CODES FROM CARDS,EACH CARD CAN HAVE UP TO 18 ENTRIES
C
      READ(5,8) CODE1,CODE2
    8 FORMAT(18F4.1)
      WRITE(6,11)
   11 FORMAT (///T56,'CODE 1',//)
      WRITE(6,9) (CODE1(I),I=1,18)
    9 FORMAT (//2X,18F6.1)
      WRITE (6,13)
   13 FORMAT (///T56,'CODE 2',//)
      WRITE(6,12) (CODE2(I),I=1,18)
   12 FORMAT (//2X,18F6.1)
      WRITE (6,106)
  106 FORMAT(///////T7,'DELT',T15,'TIMES',T27,'SIGN',T37,'DELDEG',T49,
     1'W',T59,'D',T67,'DSTEP',T76,'START',T87,'STOP',T97,'SPACE',T104,
     2'NRANTS',T112,'START-T',T122,'STOP-T',//)
C
C
C     DELT=COMPUTATION INTERVAL
C
C
      DELT=0.001
C
C
C     READ IN CONSTANTS FROM CARDS IN THE FOLLOWING ORDER,
C     SIGN,DELDEG,D,START,STOP,START-T,DSTEP,NRANTS.
C
```

```
C
      READ (5,100) SIGN,CELCEE,D,STARR,STCO,GO,CSTEP
  10C FCRMAT (F11.6)
      READ (5,105) NRANTS,JJ,JJJ,JJJJ
  1C5 FCRMAT (I11)
C
C
C     SIGN=SUMMATION SIGN FCR ARRAY,SIGN=1...ADC,SIGN=-1...SUBTRACT
C     CELDEG=SHIFT IN ANGLE OF INCOMING SIGNAL..IN CEGREES
C     C=INDEX TO VARY THE PHASE SHIFT CF THE DELAY LINES IN THE FILTER
C     IF D=1, THEN THE DELAY IN THE FILTER=THE ANTENNA SEPARATION
C     START=THE STARTING VALUE IN DEGREES OF THE INCOMING SIGNAL
C     E.G. A SIGNAL CCLINEAR WITH THE LINE OF THE ARRAY=0.0 DEGREES.
C     STOP=THE FINAL VALUE IN DEGREES CF THE INCCMING SIGNAL
C     START-T=TIME TO START COMPUTATION,E.G. T=C.C
C     CSTEP=CCMPUTATION INTERVAL FCR THE PARAMETER D.
C     JJ=CONTROL FOR PRINTING TABULATICN CF DATA(TIME,CUTPUT-A,
C     OUTPUT-B,FILTER OUTPUT....IF JJ=2..DO NOT PRINT...
C     JJJ=CONTROL FOR AUTCCORRELATICN CR CROSSCCRRELATION..IF JJJ=1 THEN
C     CROSSCORRELATE...IF NOT EQUAL TC CNE,E.G. JJJ=2 THEN AUTCCCRRELATE
C     JJJJ=CCNTRCL TO PLCT MAX FILTER CUTPUT VS MATCHING CCEFF
C     IF JJJJ=2 THEN SKIP THIS PLCT..
C     NRANTS=NUMBER OF INDIVIDUAL ANTENNAS IN EACH ROW
C
C
C     COMPUTE W,SPACE,STOP-T AND TIMES IAW CODE LENGTH.
C
C
      STCP=STOO*0.017453
      START=STARR*0.017453
      DELDEG=DELDEE*0.017453
      CM=2.0-D
      DL=D
C
C
C     RATIO OF PULSE DURATION TO ANTENNA PROPAGATION DELAY IS CETERMINED
C     BY INITIAL CHOICE CF 'SPACE' AND 'W'.  TC CHANGE THE RATIO, PROGRAM
C     CARDS MUST BE CHANGED.
C
C
      SPACE=0.1
      W=0.2
C
C
C     CHECK TC ENSURE THAT CORRELATION PROCESS CAN BE CCMPLETED IN
C     2000 COMPUTATIONS
C
C
```

```
      JHALT=1
 2222 HALT=(2.0*(SPACE*(NRANTS-1)))+W
      CHECK=HALT/DELT
      IF(CHECK.LT.1999.0) GO TO 4444
      SPACE=SPACE/2.0
      W=W/2.0
      JHALT=JHALT+1
      IF (JHALT.EQ.20) GC TO 4000
         GO TC 2222
 4444 TIMES=1.0/DELT
C
C
C     W=LENGTH (DURATION) OF THE INCOMING SIGNAL
C     SPACE=ANTENNA SEPERATION,ANTENNAS HAVE UNIFORM SPACING
C     STOP-T=TIME TC STCP CCMPUTATION,E.G. T=16.0
C     TIMES=MULTIPLYING FACTOR USED IN SUB-PROGRAMS.
C
C
      WRITE (6,200) DELT,TIMES,SIGN,DELDEE,W,D,DSTEP,STARR,STCE,SPACE,
     1NRANTS,GO,HALT
  200 FCRMAT (' ',10F10.4,I5,F13.3,F9.3,/)
      IF (JJ.EQ.2) GO TO 601
      WRITE (6,603)
  603 FCRMAT (///T20,'TABULATE CORRELATICN DATA',//)
      GC TO 605
  601 WRITE (6,602)
  602 FCRMAT (///T20,'DO NCT TABULATE CCRRELATICN DATA',//)
  605 IF (JJJ.EQ.1) GO TO 606
      WRITE (6,604)
  604 FCRMAT (///T20,'PRCCEEDING WITH ALTOCORRELATICN',//)
      GC TO 608
  606 WRITE (6,607)
  607 FCRMAT (///T20,'PROCEEDING WITH CRCSSCORRELATION UTILIZING ',
     1'CRTHOGCNAL COMPLEMENTARY SEQUENCES',//)
  608 IF (JJJJ.EQ.2) GO TC 610
      WRITE (6,609)
  609 FCRMAT (///T20,'PLCT MAXIMUM FILTER OUTPUT VS MATCHING CCEFF',//)
      GC TO 204
  610 WRITE (6,611)
  611 FCRMAT (///T20,'DO NCT PLOT MAXIMUM FILTER OUTPUT VS',
     1' MATCHING CCEFF',//)
  204 WRITE (6,205)
  205 FCRMAT (//'+++++++Y+++++++++++++++++++++++++++++++++++++++++++++++',
     1'++++++++++++++++++++++++++++++++++++++++++++++++++Y++++++++++++++',/)
      IF (NRANTS.GT.12) GC TO 210
      AAAA=24.0
      GC TO 5
  210 IF (NRANTS.GT.20) GO TO 220
```

```
      AAAA=48.0
      GC TO 5
  220 AAAA=72.0
    5 CC 4 I=1,200
      TA(I)=0.0
      TCUT(I)=0.0
    4 CCNTINUE
      IC=0
    1 ANGLE=START
      DC 16 I=1,2000
      ACD2(I)=0.C
      ACC1(I)=0.0
      ACC(I)=C.0
      PT(I)=0.0
      PCUT(I)=0.0
   16 CCNTINUE
      DO 17 I=1,400
      TIME(I)=0.0
   17 CCNTINUE
C
C
C     START COMPUTATION WITH INITIAL CR NEW VALUE OF ANGLE OF ARRIVAL
C
C
    2 DELAY=SPACE*CCS(ANGLE)
      FILTER=D*DELAY
      CUTMAX=0.0
      BRNG=ANGLE*57.2958
      T=G0
      K1=1
      KK=C
      CC 3 I=1,5C
      SIG(I)=0.0
      SUM(I)=C.0
    3 CCNTINUE
C
C
C     IC=COUNTER FOR PLOTP SUBPROGRAM...C VS MAX AMPLITUDE FOR FIXED AOA
C
C
      IC=IC+1
C
C
C     START COMPUTATION AT THE INITIAL CR NEW TIME
C     M INDICATES THE CODE IN USE,M=1 INDICATES CCDE 1 IN USE
C
C
    6 M=1
```

```
C
C
C     LCAD CCDE 1 INTO THE DUMMY ARRAY
C
C
      DC 10 I=1,18
      IDUMB(I)=CODE1(I)
   10 CCNTINUE
C
C
C     CCMPUTE SIGNAL AT OUTPUT OF EACH INDIVIDUAL ANTENNA
C
C
      I=1
   15 SIG(I)=F(T,I,W,DELAY,TIMES)
      I=I+1
      IF(I.LE.NRANTS) GO TO 15
C
C
C     CCDE ANTENNA OUTPUTS AND SUM
C
C
   19 K2=1
      ACC(K1)=IDUMB(K2)*SIG(K2)
      K2=K2+1
   20 ADD(K1)=ADD(K1)+IDUMB(K2)*SIG(K2)
      K2=K2+1
      IF (K2.LE.NRANTS) GC TC 20
      IF (M.GT.1) GO TO 21
      ADD1(K1)=ADD(K1)
      GC TO 22
   21 ADD2(K1)=ADD(K1)
C
C
C     CCMPUTE SIGNALS AT THE OUTPUT OF EACH DELAY LINE IN THE FILTER
C
C
   22 J=1
   25 SUM(J)=G(T,J,FILTER,DELT,M,TIMES)
      J=J+1
      IF(J.LE.NRANTS) GO TO 25
      IF (JJJ.NE.1) GO TO 30
      IF (M.EQ.2) GO TO 29
C
C
C     GENERATE FILTER CCDE FCR OUTPUT-A
C
C
```

```
      DO 28 I=1,18
      JDUMB(I)=-1.0*CODE2(I)
   28 CONTINUE
      GO TO 26
C
C
C     GENERATE FILTER CODE FOR OUTPUT-B
C
C
   29 DO 27 I=1,18
      JDUMB(I)=CODE1(I)
   27 CONTINUE
      GO TO 26
C
C
C     REVERSE THE CODE
C
C
   30 JK=1
      KDUMB(JK)=IDUMB(JK)
   37 JK=JK+1
      KDUMB(JK)=IDUMB(JK)
      IF (KDUMB(JK).NE.0.0) GO TO 37
      JKJ=JK-1
      DO 38 KJ=1,JKJ
      KKK=JK-KJ
      JDUMB(KJ)=KDUMB(KKK)
   38 CONTINUE
C
C
C     CODE DELAY LINE OUTPUTS AND SUM
C
C
   26 L=1
      OUTPUT=JDUMB(L)*SUM(L)
      L=L+1
   35 OUTPUT=OUTPUT+JDUMB(L)*SUM(L)
      L=L+1
      IF(L.LE.NRANTS) GO TO 35
      IF(M.GT.1) GO TO 55
      AOUT=OUTPUT
C
C
C     CALL UP CODE-2
C
C
      DO 45 I=1,18
      IDUMB(I)=CODE2(I)
```

```
   45 CCNTINUE
      M=M+1
      GC TO 19
   55 BCUT=CUTPUT
C
C
C     SUM THE FILTER CUTPUTS...THIS IS THE FINAL OUTPUT
C
C
      SIGOUT=AOUT+SIGN*BCUT
C
C
C     KK=COUNTER FOR PLOTP SUBPROGRAM...TIME VS AMPLITUDE
C
C
   56 KK=KK+1
      PT(KK)=T*TIMES
      PCUT(KK)=SIGOUT
   64 IF(OUTMAX.GE.SIGOUT) GO TO 65
      CUTMAX=SIGCUT
   65 TOUT(IC)=OUTMAX
      TA(IC)=D
      T=T+DELT
      K1=K1+1
      IF (T.LE.HALT) GO TO 6
      KK=KK+1
      PT(KK)=C.C
      PCUT(KK)=0.0
      KK=KK+1
      PT(KK)=HALT*TIMES
      PCUT(KK)=AAAA
      BBBB=ARCCS(D*COS(ANGLE))
      FILANG=BBBB/0.01745
      IF (JJJ.EQ.1) GO TO 902
      WRITE (6,901) BRNG,D,FILANG
  901 FCRMAT (///T30,'AUTOCORRELATION PLCT FOR AN ARRIVAL ANGLE ',
     1'CF',F5.1,' DEGREES',//T30,'AND A FILTER DELAY',F6.3,' TIMES THE',
     2' ANTENNA DELAY---THE',//T30,'FILTER IS MATCHEC TC AN ARRIVAL',
     3' ANGLE CF',F5.1,' CEGREES.',//)
      GC TO 904
  902 WRITE (6,903) BRNG,D,FILANG
  903 FCRMAT (///T30,'CROSSCORRELATION PLOT FOR AN ARRIVAL ANGLE ',
     1'CF',F5.1,' CEGREES',//T30,'AND A FILTER CELAY',F6.3,' TIMES THE',
     2' ANTENNA DELAY---THE',//T30,'FILTER IS MATCHED TC AN ARRIVAL',
     3' ANGLE CF',F5.1,' CEGREES.'//)
  904 IF (JJ.EQ.2) GO TC 70
      WRITE (6,1000)
 1000 FCRMAT (//T12,'TIME',T24,'OUTPUT-A',T40,'OUTPUT-B',T52,
```

```
     1'FILTER OUTPUT',//)
      WRITE (6,2000) T,ACUT,BCUT,SIGOUT
 2000 FORMAT (' ',4F15.4,/)
   70 CALL PLOTP(PT,POUT,KK,0)
 2800 FORMAT (//T30,'TIME (HORIZ) UNITS ARE NORMALIZED TO INPUT PULSE',
     1' WIDTH--',//T30,'200 UNITS=ONE PULSE.  AMPLITUDE (VERT) UNITS',
     2' ARE NORMALIZED',//T30,'TO PULSE HEIGHT.',//)
   67 WRITE (6,2300)
 2300 FORMAT (//'*********************Y*******************************************************',
     1'***********************************************************Y***************',/)
      LAST=0
      CALL DRAW(900,PT,POUT,0,0,LABEL,TITLE,200.0,5.0,3,2,2,0,5,6,1,
     1LAST)
C
C
C     INCREMENT THE FILTER DELAY BY DSTEP
C
C
 3600 D=D+DSTEP
      IF (D.LE.DM) GO TO 1
 2950 IF (JJJJ.EQ.2) GO TO 3700
C
C
C     PLOT MAXIMUM FILTER OUTPUT VS MATCHING COEFF FOR A FIXED ACA
C
C
      WRITE (6,3001) BRNG
 3001 FORMAT (///T30,'PLOT OF MAXIMUM FILTER OUTPUT VS MATCHING COEFF',
     1' (D)',//T30,'FOR A FIXED ARRIVAL ANGLE OF',F5.1,' DEGREES'//)
      IC=IC+1
      TA(IC)=DU
      TOUT(IC)=0.0
      IC=IC+1
      TA(IC)=DM
      TOUT(IC)=AAAA
      WRITE (6,3500)
 3500 FORMAT ('0','0',/)
      CALL PLOTP(TA,TOUT,IC,0)
      WRITE (6,3000)
 3000 FORMAT (//T30,'THE MATCHING COEFFICIENT (D) IS PLOTTED ALONG',//
     1T30,'THE X-AXIS AND THE MAXIMUM FILTER OUTPUT IS PLOTTED',//
     2T30,'ALONG THE Y-AXIS.'//)
C
C
C     INCREMENT THE ANGLE BY DELDEG
C
C
 3700 ANGLE=ANGLE+DELDEG
```

```
      IF (ANGLE.LE.STOP) GO TO 2
      GO TO 6000
 4000 WRITE(6,5000)
 5000 FORMAT(///T20,'DATA EXCEEDS ARRAY STORAGE...JOB TERMINATED',///)
 6000 STOP
      END

      FUNCTION F(T,I,W,DELAY,TIMES)
      INTEGER*4 T3,FIL3,DELT3,W3,D3
      T1=T
      T2=TIMES*T1+0.1
      T3=IFIX(T2)
      W1=W
      W2=TIMES*W1+0.1
      W3=IFIX(W2)
      D1=DELAY
      D2=TIMES*D1+0.1
      D3=IFIX(D2)
C
C
C     SNR AT ANTENNA OUTPUT IS DETERMINED BY VALUES OF 'F'...SIGNAL IS
C     NORMALIZED AT F=1.0, NOISE CAN BE VARIED FROM 0.0 TO 1.0 BY
C     CHANGING PROGRAM CARDS.
C
C
      F=0.0
      IF(T3.LT.((I-1)*D3)) GO TO 190
      F=1.0
      IF(T3.LE.(((I-1)*D3)+W3)) GO TO 190
      F=0.0
  190 RETURN
      END

      FUNCTION G(T,J,FILTER,DELT,M,TIMES)
      COMMON/ZZZ/ADD(2000)
      COMMON/AAA/ADD1(2000)
      COMMON/BBB/ADD2(2000)
      INTEGER*4 T3,FIL3,DELT3,W3,D3
      T1=T
      T2=TIMES*T1+0.1
      T3=IFIX(T2)
      FIL1=FILTER
      FIL2=TIMES*FIL1+0.1
      FIL3=IFIX(FIL2)
      DELT1=DELT
      DELT2=TIMES*DELT1+0.1
```

```
      DELT3=IFIX(DELT2)
      G=C.0
      IF(T3.LT.((J-1)*FIL3)) GO TO 290
      KKK=((T3-((J-1)*FIL3))/DELT3)+1
      IF (M.GT.1) GO TO 291
      G=ADD1(KKK)
      GO TO 290
  291 G=ADD2(KKK)
  290 RETURN
       END
```

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Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

...GINATING ACTIVITY (Corporate author) Naval Postgraduate School Monterey, California 93940	2a. REPORT SECURITY CLASSIFICATION Unclassified
	2b. GROUP

...ORT TITLE

Technical Evaluation of a Radio Direction Finding System Utilizing Complementary Sequence Phase Coding of Incident Signals

...SCRIPTIVE NOTES (Type of report and, inclusive dates)

Master's Thesis; June 1973

...THOR(S) (First name, middle initial, last name)

David Louis Prisaznick

...PORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
June 1973	179	7

...NTRACT OR GRANT NO.	9a. ORIGINATOR'S REPORT NUMBER(S)
...OJECT NO.	9b. OTHER REPORT NO(S) (Any other numbers that may be assigned this report)

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...STRACT

A new radio direction finding system has been proposed which offers advantages over current systems in the areas of bearing accuracy and simplicity of antenna design requirements. This system phase codes the outputs of a linear array of omni-directional antennas, and then combines these time-separated signals so that the sequences produced uniquely determine the signal arrival angle. A correlation process using matched filters is performed on the resulting sequences in order to determine the corresponding arrival angle.

Resolution of the inherent ambiguities of the system, due to elevation angle and adjacent quadrant azimuth components is demonstrated using a two-array technique. A computer simulation of the phase coding/decoding process, using an idealized signal model is used to evaluate the performance of the system for various input noise conditions, array lengths, and input signal parameters.

KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
radio direction finder						
phase coding						
complementary sequences						
time of arrival						
interferometer						